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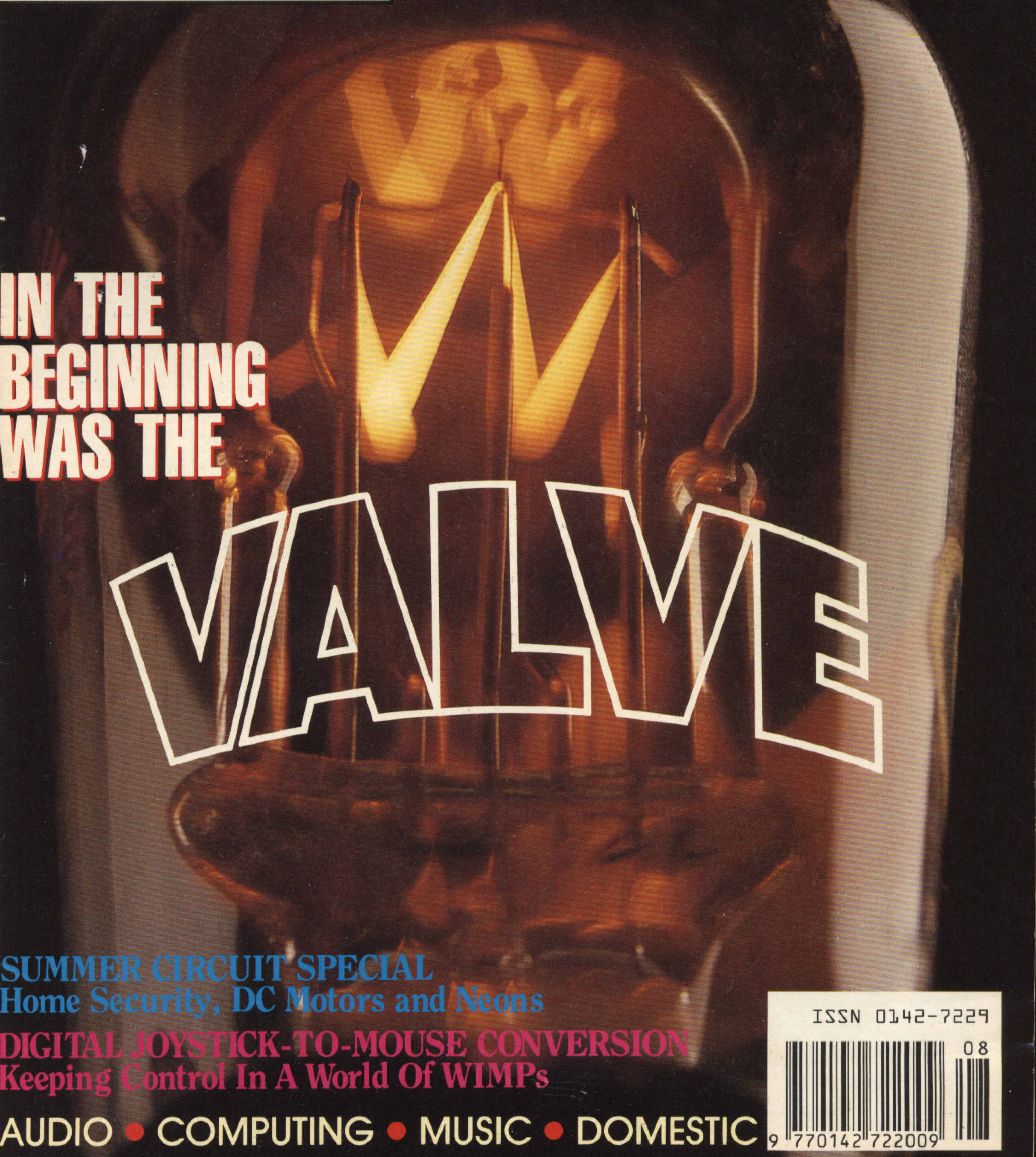

IN THE BEGINNING WAS THE VALVE

SUMMER CIRCUIT SPECIAL
Home Security, DC Motors and Neons

DIGITAL JOYSTICK-TO-MOUSE CONVERSION
Keeping Control In A World Of WIMPs

AUDIO • COMPUTING • MUSIC • DOMESTIC

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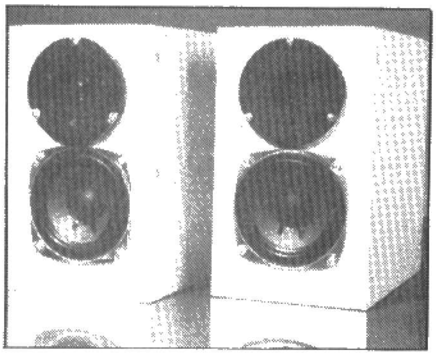

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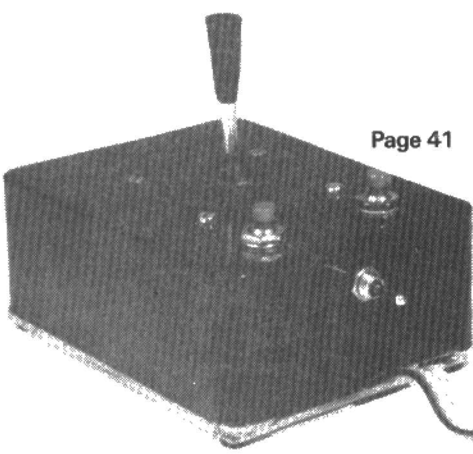

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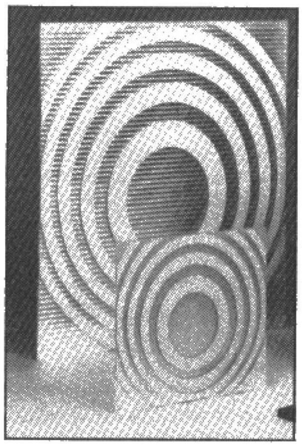
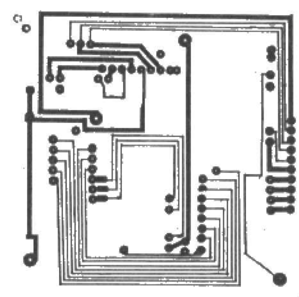
REGULARS





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ETI
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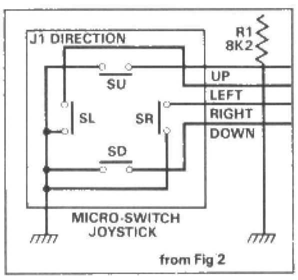
TODAY INTERNATIONAL

FEATURES/PROJECTS



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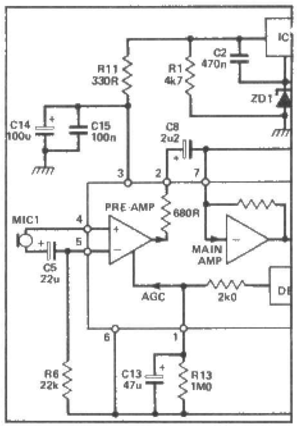

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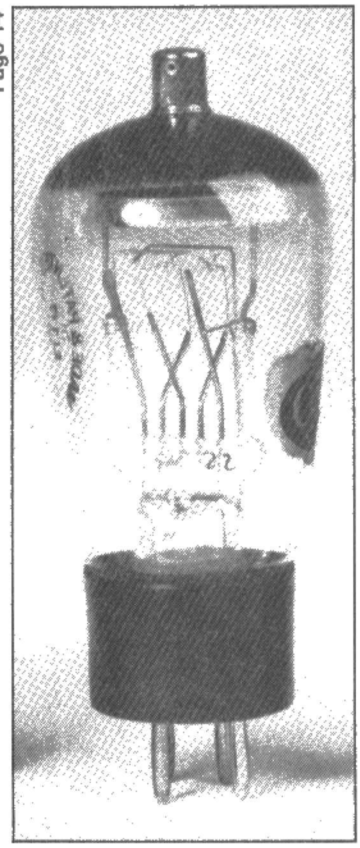

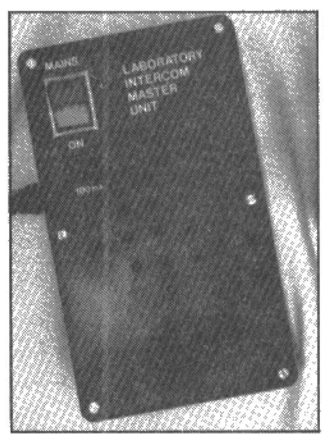

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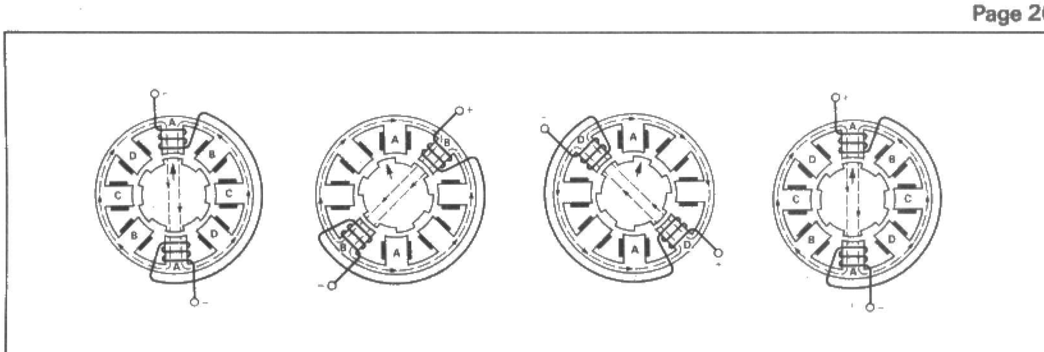
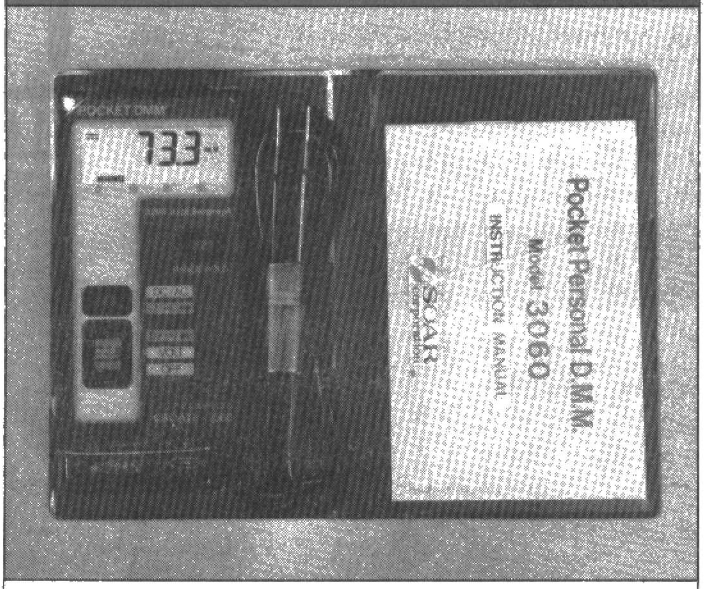


POCKET DMM



Solex International has a range of pocket sized digital multimeters. The 3060 model is a 3½ digit display and measures in at 51×106×10mm. It offers high speed auto or manual randing, measuring AC or DC voltages up to 450V and resistances from 300R to 30M.

The pocket sized DMM also tests diodes and has a useful continuity bleeper function. It delivers high speed sampling on a 32-segment bar graph display.

Solex claims it is electrically protected on all ranges and is virtually 'drop proof'.

A good feature about this meter is the atomatic power off. How many times have you left portable equipment on, only to find exhausted batteries on return!

The DMM comes with instruction manual, test leads, protective case and two batteries.

The 3020 model has similar features except for a maximum resistance range of 2M0 and a 32-segment bar graph display.

The 3060 retails at £29 plus VAT.

Further details from Solex International, 95 Main Street, Broughton Astley, Leicestershire LE9 6RE.

CD-I

Sony, Matsushita and Philips have announced that they will be joining forces for the promotion and marketing of the latest CD technology, Compact Disc Interactive (CD-I).

Compact Disc Interactive combines high quality sound, text, still image and full motion picture, computer graphics and data all on a 5-inch disc. All these forms of information can be used interactively via a keyboard or mouse control.

The three companies will be working together on improving the technical capabilities of the CD-I system, particularly in the field of high quality video. High definition TV is not far away!

This powerful machine could well find its way into many homes for use as an entertaining medium, but perhaps more importantly in supplying a wide variety of local and national information.

Another area of use will be in education, where discs on various individual topics could be used at schools and colleges or indeed for home tutorials. Several companies are in the process of preparing CD-I titles.

QUANTUM LEAP FOR BELL

AT&T Bell Laboratories, the people that brought you the first transistor, have demonstrated a new quantum device that functions as a tiny parallel processor to reduce circuit complexity dramatically.

The device is called a multi-state resonant tunnelling bipolar transistor no less!

Until now, normal transistors have been limited to two states: 'on' and 'off'. This new transistor is the first with multiple states: three or more.

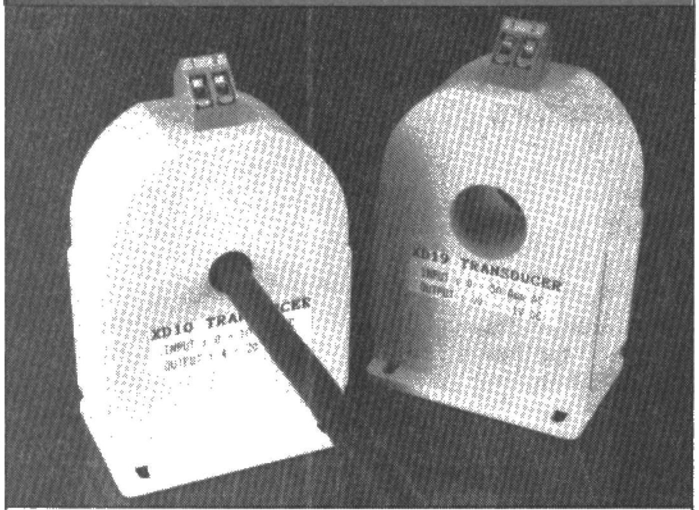
When the input current increases in the device, the output current peaks, falls off, then peaks again. This unique, multi-peak characteristic produces the multiple states and allows a single device to do the work of many conventional transistors.

An example is in parity bit checking in digital systems. They normally use 24 conventional transistors to check four bit words. The scientists at Bell have achieved this task using just one of the new devices.

The demonstration suggests that in future a much higher density of operations could be packed in the same area as today's integrated circuits.

Although this is a great advance, resonant tunnelling transistors are still a research technology and many development hurdles remain.

CURRENT TRANSDUCERS



From Northern Design comes a selection of novel current transducers that are designed for both instrument and data logging applications.

These new units can detect from 5 to 300A through the wire they surround. They can have a DC output of 1V and be self powered or a 4 to 20mA current output which requires an external 24V DC supply. Operating frequency is between 40Hz-5kHz and source impedance lower than 500R. Prices are from £33 plus VAT.

Contact Northern Design (Electronics) Ltd, 228 Bolton Road, Bradford, West Yorkshire BD3 0QW. Tel: (0274) 729533.

LAG TAG ON TRIAL

An experimental system of electronic tagging for convicted criminals will soon commence using 'selected persons'.

The Government intends to reduce the bulging prison population by this new electronic tracking method. The offender would have to wear a small waterproof, calculator-sized monitor around the neck or under their clothes. The device would emit coded signals three times an hour to the nearest specialised cellular radio station, which then passes the signal on to a central computer. The 'cells' would cover areas ranging up to two miles square.

The tag wearer would be restricted to a certain number of cells for normal movements including access to work. Should the wearer move out of the permitted cells, the monitor would emit audible signals and also inform the central computer to alert the authorities.

A trial has recently been completed at the University of Kent using student volunteers.

CHEAPER APRICOTS FOR SALE

Apricot Computers of Birmingham, who specialise in desktop computers, are producing their first generation of powerful machines for large businesses.

Processing power is available for up to 128 workstations and is likely to make use of the latest 80/486 microprocessor from Intel (*see Insight*).

This 32-bit processing machine should provide stiff competition against other computing systems costing between four and twenty five times the price.

FREE CAT

Falcon Electronics has its latest catalogue (PL18) available. It contains a large range of speaker kits for the audiophile plus capacitors, connectors and audio publications.

Send SAE with a 26p stamp for their latest list to: Falcon Electronics, Tabor House, Norwich Road, Mulbarton, Nr Norwich, Norfolk NR14 8JT. Tel: (0508) 78272.

CLASSY CANS

Amongst the many pairs of headphones Sony sells there are two rather interesting pairs on offer. The first, the MDR-R10, boasts of producing concert hall quality acoustics and ergonomic comfort as never before. Have we not heard this before? Yes, well read on.

"The MDR-R10 employs a newly developed bio-cellulose diaphragm, which has over ten times the rigidity of a conventional paper cone, yet achieves naturalness and remarkable presence."

OK, a very thin tough diaphragm, that's an improvement. The earphone housing is made of 200 year old zelkova wood. This is used for its 'lightweight and solid character providing unequalled acoustic quality'. We must be approaching the law of diminishing returns here. That goes for the wood as well. But wait, there is more. To ensure luxurious comfort, the earpads are made of Greek sheepskin and the cord is wrapped in pure silk! This could get very sensuous. They are also kind enough to throw in a rhodium-base pure gold plated plug and a hard case with lock. It could really do with an anti-theft device because here is the crowning glory — the price. A mere snip at £2,499.95.

Sony does have a pair of headphones, model number MDR-1F5K, which says goodbye to the cord that invariably strangles you, called a wireless headphone. Now you can have the freedom to wander around the living room listening in digital audio quality. The system adopts an infra-red transmitter to sit on your hi-fi giving a wide dynamic transmission range of 90dB. This ensures excellent sound reproduction from compact disc. The unit's automatic level control circuit also allows for optimum FM modulation without any adjustment to provide the best listening at all times. The headphones have built-in rechargeable batteries that can be revived by placing the headphones in a stand attached to the transmitter. The frequency response of the whole system is 18Hz to 22kHz. The cost of the complete system is £99.95.

FLAT AERIAL MAT FOR SAT

Amid the chaos of melting Astra dishes and failing squarials, an entirely new range of receivers has been produced by Mawzones of Baldock in Herts.

For some time Mawzones has been producing window-mounting sheets that focus received satellite signals to a feeder behind the window (see *News* ETI Dec 1988). The focussing is performed by a set of concentric circles that are opaque to electro-magnetic radiation. The alternating transmitting and opaque areas act as Fresnel zones and focus the signals. Since the window surface is vertical and the satellite source is elevated, the circular zones are stretched to ellipses.

However the window-mounting Mawzone sheets lost half the signal strength in the opaque area and were limited in size by available window space. As a result they were best used only for low and medium rate data reception.

The new range of antennae has the transparent sheet combined with a reflective surface, the two plates being separated by a gap of a few millimetres. This effectively creates two separate antennae with a common focus in front of the sheets. The plates are mounted flush against a wall facing roughly south, with a standard low noise blockdown converter (LNB) feeding a receiver inside.

The mirror sheet designed to receive Astra is just slightly larger than its dish equivalent and will be circular. Its flush fitting makes it less vulnerable to weather damage or vandalism. Mawzones claims it can be painted without affecting the output, thus making it less of an eyesore (the left of the larger plates in the photo is a Eutelsat receiver disguised as roof slates).

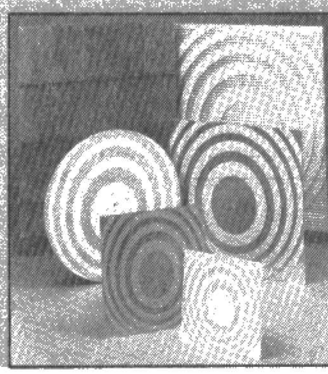
The fact that the plates are flush fitting means that even the larger dishes are exempt from planning permission.

A target price has not yet been set for the Astra antennae but it will be aimed around the same level as conventional dishes. Mawzones sees "no reason to undercut conventional dishes since most people, given the choice of two similarly priced antennae, would opt for obvious advantages of the Mawzones design".

A design for receiving BSB is already well under way. When BSB's problems with the squarial became apparent, Mawzones suggested to BSB that the sheet antennae could solve its problems. Reorientating the square sheet antennae as a diamond could even save face for BSB if it had to dump the squarial. A dummy BSB design was prepared by Mawzones but BSB seems intent to continue with its delayed squarial plans.

Mawzones has also developed an enormous sheet antenna (4m × 4m) capable of use with C band transmissions. The sheet is modular, consisting of many tiles each with part of the concentric circles pattern so that the antenna can be easily transported and erected in remote areas.

Contact Mawzones Ltd, 6 Hodwell, Ashwell, Baldock, Herts SG7 5QG. Tel: (0462) 742854.



SIGNAL GETS ALL CLEAR

A team at the British Telecom Research Laboratories has developed an optical amplifier to boost weak optical signals along their newer communication routes.

This is another major step forward because at present all optical signals are converted to electronic ones which are then amplified in the normal manner, only to be reconverted to optical once again to continue on their route.

Apart from overcoming this disadvantage, this latest amplifier will work at many different wavelengths and bit rates. It is not affected by the polarisation of the input signal, something that earlier amplifiers suffered from.

"Fault location in lines will now be made much easier with this new device because with the existing system any optical signal designed to find a fault is interrupted by the electronic repeater."

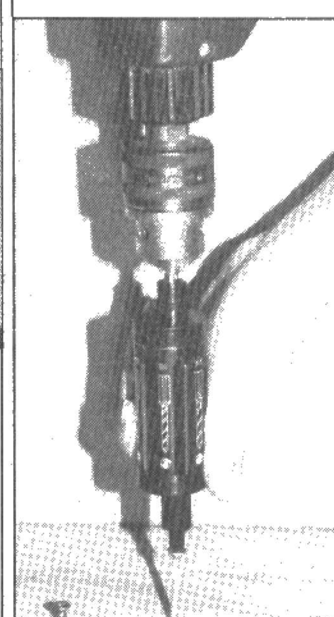
DIAL A DRILL

Now available from Freetrade (TEP) Ltd., is the new Dial A Drill conversion unit.

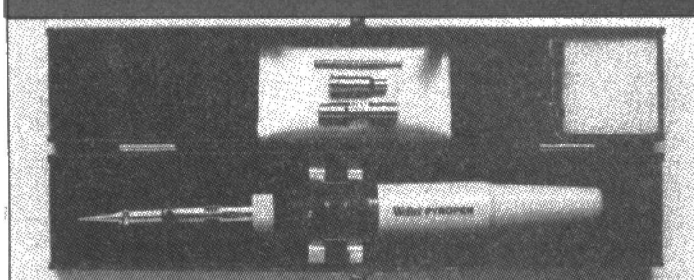
This unit will convert any power drill into a six-piece self-contained screwdriver set. The unit contains four drill bits, 1/16 in, 1/8in, 1/16 in, 1/4in and two screw drivers, No. 2 Phillips and 1/4in slotted. All are of high quality and heat treated for strength, including HSS treatment.

For ease of operation the desired bit can be selected quickly and simply.

Contact Freetrade (TEP) Ltd., Unit 15C, Aviary Industrial Park, Garrison Lane, Bordersley Green, Birmingham B9 4QE. Tel: (021) 766 6142.



PORTABLE PYROPEN



From Axiom Electronics comes a cordless soldering iron, made by Weller.

The soldering iron is ideal for use in field service applications. Pyropen is an LP gas-operated iron which is independent of mains or batteries and weighs only 90g. The gas in a single charge is capable of giving four hours continuous operation.

The kit also includes a 3.3mm tip, a 5.7mm hot blow tip, torch ejector, spanner, cleaning sponge and carrying case.

The price is £54.08 plus VAT.

Contact Axiom Electronics on (0494) 461616.

BLUEPRINT

This column is a service to readers to provide electronic designs. Send your requirements, with as much detail as possible, to ETI Blueprint, Argus House, Boundary Way, Hemel Hempstead HP2 7ST.

This month the question is from Peter Morgan in the county of Powys. He would like a design for a bargraph display of audio spectral content. He requests displays of the level at 100Hz, 200Hz, 400Hz, 600Hz, 800Hz, 1kHz, 4kHz, 6kHz, 8kHz and 10kHz, and says that it is intended to be plugged into a headphone socket.

For an indicator of this type to be meaningful, it must have equal proportional intervals between frequencies in order to correspond approximately with the way in which sound is perceived. 2:1 frequency steps would be reasonable, so to fit ten frequencies into the audio range we would aim for frequencies of: 32Hz, 64Hz, 128Hz, 256Hz, 512Hz, 1024Hz, 2048Hz, 4096Hz, 8192Hz and 16384Hz. In practice, variations of up to 20% either side of these frequencies are unlikely to affect the usefulness of the display.

Building Blocks

The sound must be filtered into frequency bands, the sound level at these bands must be detected and the result must be displayed on LED bargraphs. A logarithmic display is necessary to cope with the logarithmic perception of sound intensity, so a logarithmic bargraph IC such as the LM3915 is the obvious answer. The need to connect the circuit to a headphone socket, with the inevitable variation of signal level as a volume control is set to different levels, necessitates a gain control at the input of the circuit. Putting all this together gives rise to the block diagram shown in Fig. 1.

The whole system, with two 10-channel indicators (one for each stereo side) encompasses a lot of circuitry but consists of a few simple building blocks repeated with minor variations. The first block, shown in Fig. 2, is the input buffer, one for each channel. This is there to provide a low impedance drive to the filter stages. It is preceded by a gain control to set the average level of the display to a reasonable level at whatever volume setting is in use on the amplifier.

Following the DC blocking capacity is a protection circuit to prevent damage to the op-amp input if the volume is turned up too high. After the unity gain buffer are ten bandpass filters, giving a total loading on the buffer of 330R.

The filters are of the multiple feedback configuration shown in Fig. 3. The configuration remains the same at all frequencies but the capacitor values change to set the centre frequency. The Q of the filter, the centre frequency divided by the bandwidth, is set by the resistor ratio and is thus constant for all frequencies. This is exactly what is required — the same proportional bandwidth because the proportional frequency spacing of the filters is the same.

A Q value between one and two is reasonable for the application, so a convenient set of resistor values is chosen to giving a Q value of 1.6. The circuit also has a voltage gain dependant on the Q value chosen. In this case the circuit gain is five, which is not too high to be practical. It does mean that the gain-bandwidth product of the op-amp used becomes significant at the higher frequencies. The LM324 would be suitable for the lower frequencies but for the top two bands a higher bandwidth type such as a TL084 would be better. On the other hand, the effect on the indication level will be minor.

The table of component values and frequencies in Fig. 3 shows the preferred value components which should be used and the actual centre frequencies which these give. If more accuracy is needed, parallel capacitor combinations could be used to give the actual values calculated from the formula. This is only worthwhile, of course, if the tolerance of the capacitors is commensurate with the accuracy required.

Level Display

To provide a useful display of the signal level within a frequency range, it is necessary to rectify it and detect the resulting DC level with suitable charge and discharge time constants. The modified precision rectifier shown in Fig. 4 represents one approach to the problem. The charge time constant is R1×C1 and the discharge time constant is R2×C1. This latter must be much longer than the charge time constant because the value of R2 must be greater than that of R1 to permit the capacitor to reach its required voltage.

This detector will therefore detect peak signal levels of negative polarity. Suitable choice of R1 and R2 could produce an approximation to a peak programme meter characteristic. Equally, choice of component values can produce a peak reading display which will fall only slowly when the signal level drops, or something like a VU meter which responds rapidly to falling signal level. The limit on the speed of response is that the discharge time constant must not allow significant ripple on the capacitor at the centre frequency of the filter or else the display will flicker. For the 32Hz channel, a time constant of about 300ms is required. Example component values for this would be: R1=1k, C1=47μ, R2=56k. For higher frequencies if a fast response is required, the value of C1 can be reduced to 10μ and the resistor values chosen accordingly.

If both halves of the cycle must be measured, the circuit of Fig. 5 would be

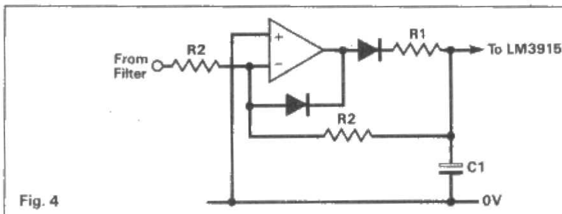


Fig. 4

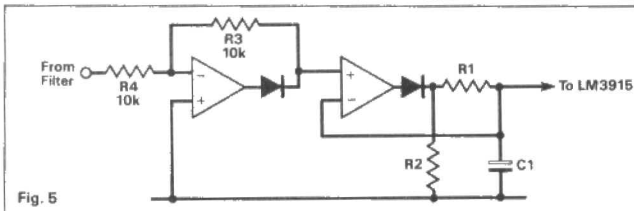


Fig. 5

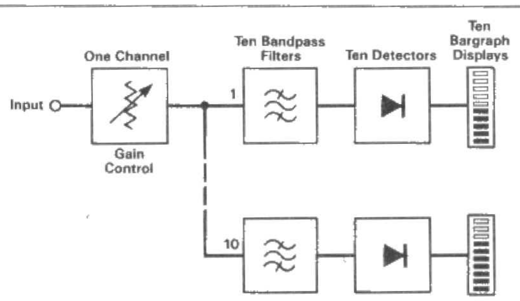


Fig. 1

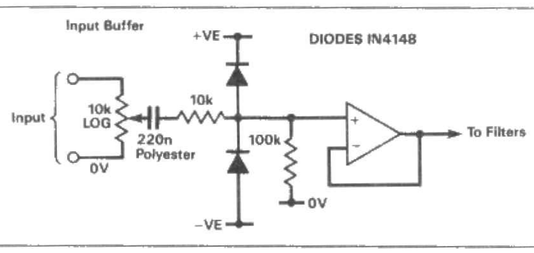
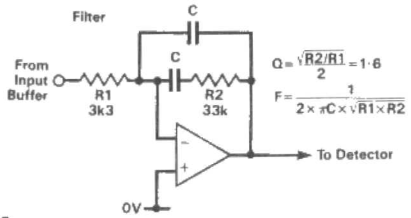


Fig. 2



$$Q = \frac{\sqrt{R2/R1}}{2} = 1{\cdot}6$$

$$F = \frac{1}{2 \times \pi C \times \sqrt{R1 \times R2}}$$

C	Frequency
470n	32.5Hz
220n	69.3Hz
100n	153Hz
56n	272Hz
27n	565Hz
15n	1016Hz
6n8	2242Hz
3n9	1910Hz
1n8	8472Hz
1n	15251Hz

Fig. 3

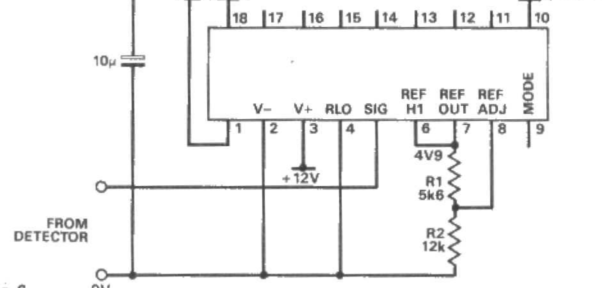


Fig. 6

suitable. This circuit can be configured to produce similar attack and decay times if R2 is significantly lower than R1. The indication will then be based on an average reading of signal level rather than on peak.

Figure 6 shows the suggested circuit to use the LM3915 bargraph display IC. As detailed later, the LED power supply is separate from the main power supply.

To provide reasonable brightness, a LED current of 5mA is chosen. The LM3915 regulates this current itself, as approximately ten times the total current drawn from the reference output (pin 7). The basis reference voltage is 1.25V (between Ref out and Ref adj) and the resistance of the resistive divider chain in the LM3915 is 22k, and the Ref adjust ground current is 80μA, so the formulae for reference voltage and LED current are

$$V_{ref} = 1.2V \times (1 + \frac{R2}{R1}) + R2 \times 80\mu A$$

$$I_{LED} = \frac{12.5V}{R1} + \frac{V_{ref}}{2k2}$$

The decoupling capacitor must be close to the LM3915 and each IC or pair of ICs should have one, depending on layout, otherwise the circuit is likely to oscillate. Adequate decoupling should also be provided on the ±12V power supplies to avoid interaction between stages. Finally the 0V tracks from the LM3915s should be starred with fairly thick wire back to the power supply, to prevent the heavy LED currents from imposing significant earth voltages relative to the signals.

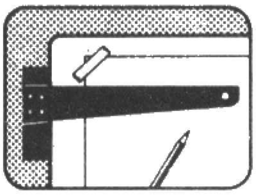
Power Supplies

The circuit should be powered from ±12V to 15V. In order to prevent the LM3915 ICs from over-dissipating, a lower voltage supply is needed for these chips. To drive a total of 20 bargraph displays each with ten LEDs and to produce a reasonable brightness requires 1A. The best way to provide this is to use a metal can (TO3) LM317 regulator on a big heatsink to supply the LEDs and to set its output voltage between 3.5-4V. This will minimise dissipation in the LM3915s, at the expense of dissipating a lot of power in the regulator.

Because of the three power supply rails required, it is probably worthwhile to use two separate mains transformers, one rated at 15-0-15V for the ±12V supplies and another rated at 9V to supply the LEDs. The 15V transformer need only be a 6VA type but the transformer powering the LEDs needs to be rated at 12VA.

With all this LED power available, this unit when completed should light up the room nicely.

Andrew Armstrong



INSIGHT

Ever since Intel launched the 33MHz 80386 processor, power users have been looking to the next generation of chips — the 486 family — for their processing panacea. But will the 486 live up to expectations?

The microprocessor world seems to be divided into two schools of thought. One side has opted for the fast but relatively untried "new" technology of the RISC (Reduced Instruction Set Computer) chip sets such as IBM's RT series and DEC's MIPS co-processors — the key to their open systems architecture and Ultrix developments in the future.

The other side has opted for the faster and more tried-and-tested grounds of the CISC processor (Complex Instruction Set Computer), whereby more instructions are packed into ever decreasing areas of silicon.

Of the most popular processors, the most likely to take on the sub-RISC achitecture (such as ARM from Acorn) is the Intel 80386 but, even though new machines from companies such as Tandon, Acer and AMP have proven to be fast and efficient in their own respective ways, the Intel 486 series seems to offer so much more.

The original prototypes of the 80386 were clocked at 12MHz — rather a conservative speed and one that was matched easily (and surpassed by unconventional wiring) by the older and more established 80286 series of processors. Intel like to take their processors up in speeds of 25% at a time, so that a floor speed of 12MHz was untenable for the short period of its life. This led to Intel boosting the power of the basic processor to 16MHz and dropping the 12MHz model completely.

Not that it mattered much anyway, since shortly after that Intel upped the specs again and released the 20MHz 386. Quite recently the 25MHz chip sets have appeared and are now cheap enough to be considered worthy of the Taiwanese clones from the Middle and Far East.

However, the big break came when the 33MHz machines were announced. The design specs (but not the processor chip sets) were distributed to a few major companies and an embargo was set to last until 10th April 1989 so that the processor could be launched with a fleet of high-powered compatible machines to follow literally minutes after the official launch.

This was, as they say, not strictly adhered to, although the general impression is that Intel got their new 33MHz chips out — and the dealers got their machines out — all claiming to be the fastest PCs in the world.

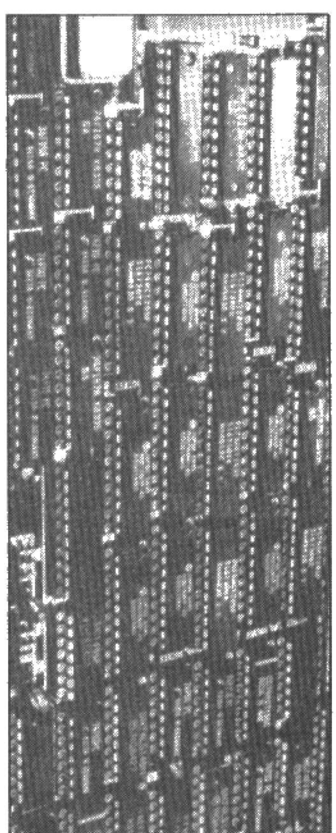
Life isn't as easy as sticking in a new chip set. Intel is well aware that their new chip needed support to properly harness the sixty per cent speed increase over a popular chip set such as the 20MHz 386. Intel has to make sure that the chip had no added features.

Bugs in short

Why all the fuss about getting the chips right in the first place? Why not wind it up and let it go — after all, there were no new fabrication processes involved. The technology — whilst operating at a slightly higher density — was well within the limits of fabrication standards. Instead Intel has refined its screening process and selected only the chips that can run at a floor level of 20MHz.

Temperature is a deciding factor — as the clock speed goes up so the number of operations goes up, requiring a larger power drain etc etc. Beyond a certain level, the chip can literally fry itself out so that gates start to short out (and not close). Errors start to occur — and as we are well aware, processor errors are always fatal ones.

So we have ascertained that speed is important and that all Intel really has done is raise its selection bandwidth — fine but what about the projected 486 chip sets — what are they and what will they run?



The 486 will be a new beginning (many people have been led to believe) offering distributed processing facilities that will be transputer-like in their ability to be linked from one machine to another. It seems likely that an entirely new tack is needed, instead of just raising the selection platform even higher (so that processors are being driven even harder). These days, the cooling systems installed inside the latest state-of-the-art microprocessor base subsystems vibrate at a level somewhat akin to a 747 taking off in your own room. I recently saw my first water-cooled PC subsystem!

These are all based on top-end 386 systems requiring no less than 16-32Mb and offering disc storage of no less than 150Mb, normally specified for power-users and emulators (running CICS IMS, Unix and other open systems ranging from IBM's SNA and SAA to the old 3083 and 370 system software — going back to CICS again!). It seems that the new family of bona-fide 32-bit bus processors are happily running more of the same software — but faster,

Putting it bluntly, the 486 will have a hard time breaking out of the mould set up by its siblings the 286 and the 386. The industry pundits and gurus have been commenting on an early 1990 production date for 486 processors. The 486 is expected to be launched with the final incarnation of the 386 chip — although it is unclear as to whether the chip will simply be a souped up mega cache version of the 33MHz 386 with built-in co-processors, or whether it will be a simple enhancement.

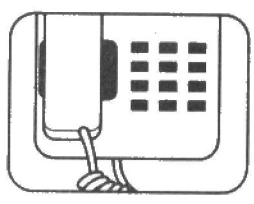
Where does that leave the PCs of the world?

Well, MCA is fast becoming the de-facto standard although the alternative EISA seems to be putting up a good fight. Indeed, many companies are placing their products directly across the two platforms (with the notable exception of Compaq). Mission's MCA portable clearly needs (or at least is happy with) as much processing power as it can get, and of course the faster processors get, the faster RAM has to get or we have to start introducing wait states — an undesirable situation.

For the time being, power users have the 33MHz chip — if they want faster they'll have to go to Digital's MIPS systems (and their nod in the direction of the Open Systems Architecture). That involves leaving the sanctity of DOS and expanded memory and that involves breaking into new ground. But then that's just what Intel may be doing in the near future anyway.

Clive Grace

OPEN CHANNEL



As business relies more and more on computers to exchange information as a faster method than old-fashioned letters-and-stamps mail, users are beginning to become seriously concerned about security. When Royal Mail takes charge of a letter, users have a high degree of security — it is, after all, illegal to open anyone else's mail. There are increased levels of mail security, *registered mail* or *Datapost*, which guarantee secure delivery — as long as the envelope is intact, the content has not been seen by anyone between sender and recipient. Further, industrial espionage agents find it extremely difficult, nay — impossible, to locate any given item of mail among the many millions posted each day. And all for 19p!

But on a computer information network it only takes a competent hacker to find required information. True, there can be codes, authorisation numbers and so on built into the system, but the more complicated the network, the cleverer and more well equipped is the hacker. It seems to be that if data is on a computer network, little or nothing will stop it being illegally removed if someone wants it bad enough. Industrial espionage is not the world of James Bond, more the concern of the computer-buff.

Hacking isn't the only concern either. Recent viruses have literally swept through some computer networks, virtually closing down networks and losing priceless amounts of files.

In the light of this, it's not really surprising that a growing number of large multinational organizations (28 at the time of writing) have joined forces in the European Security Forum. The forum is being organised by the European wing of the world-wide accounting and financial services company Coopers and Lybrand, which last year produced a report funded by the European Commission on network security.

Membership so far includes such big guns as British Airways, British Telecom, Volvo, Olivetti, Digital Equipment Corporation and the Society for Worldwide Interbank Financial Telecommunications (SWIFT) — to name but a few. A similar group called 1-4 was set up in the United States three years ago.

United We Stand

Meanwhile, the European Commission is trying to get a pan-European electronic data interchange (EDI) standard up and running as part of the committment to harmonisation by 1992. The aim is to allow networks to communicate such that electronic trading can take place between countries. To do this the Commission is funding and organising a project called EDI/Unite. Looks like the European Security Forum has been set up at just the right time. Could someone introduce the two groups?

EFTPOS

Electronic Funds Transfer at Point Of Sale, the dream of large shopping concerns, is to get a big boost this summer when shops in three selected cities (Edinburgh, Leeds, and Southampton) are fully kitted out with point of sale terminals capable of doing the biz. At its fullest, shoppers will simply have to pass a card through a card reader at a terminal, key in a personal identification number (PIN) and the goods have been paid for. All this is possible because the terminal links up to bank networks to locate the shoppers' bank accounts, automatically debiting them and simultaneously crediting the shop's bank account.

The three-city trial is being setup by EFTPOS UK, an organisation created by the banks' joint clearing house system and, despite disagreement and similar proposals by individual banks, looks set to become the first stage in a countrywide coordinated network.

Of course with such a system not only the large network users need to be concerned about network security. The possibility of illegal funds transfer exists if the network is not secure. And the only way you'll ever know it has occurred is when your bank statement drops on your doormat.

BT Moves to TV

British Telecom has always stated it could not justify the use of optical fibre in the local loops between local exchanges and subscribers homes merely for telephone purposes. The reason is financial — the cost of installation can never be covered by line charges when the line is used just for telephone calls. (Interestingly, other companies argue that optical fibre *is* justifiable, but that's another matter!)

In retrospect, BT's argument has been a good manoeuvre. Current regulations prevent BT from distributing television transmissions over the telephone network. On the other hand, cable television distribution companies *can* carry telephone calls. BT has always thought this slightly unfair. By forcing the Department of Trade and Industry's hand with its no-TV no-optical-fibre stance, BT looks as though it is going to gets its way in a change of regulations to allow television distribution over the telephone network.

If this happens, you'll be surprised how quickly your local line will change to optical fibre!

Recently BT has moved to bypass the no-TV regulation in two ways. First, a special trial (granted by the DTI) is about to start in Bishops Stortford (scheduled for March of next year, to be precise). The trial will involve some 500 users, both business and residential, providing stereo television and radio, data communications and, of course, telephone. It's to be a two-year trial costing around £5 million. Two different systems will be tested during the trial period.

Second, in a crafty move, BT's own cable television franchise subsidiary (yes, a loop-hole in the regulations) Cable Thames Valley, will soon carry telephone traffic over its optical fibre network as a so-called trial. Telecoms watchdog, Oftel, knows about the 'trial' and will be looking at it carefully. Around 200,000 homes in the High Wycombe, Newbury and Reading area will be networked and cabling should be starting any time now.

Keith Brindley

DIARY

EWEC 89 — July 10-13th
Wind energy conference and exhibition. Scottish Conference and Exhibition Centre, Glasgow.

Holographic Systems, Components And Applications — July 11-13th September
University of Bath. Second International Conference. Contact IEE on 01-240 1871

Image Processing And Its Applications — July 18-20th
University of Warwick. Third International Conference. Contact IEE on 01-240 1871

Vacuum Microelectronics — July 24-26th
University of Bath. Conference sponsored by The Institute of Physics, IEE and IEEE. Contact The Institute of Physics on 01-235 6111

Eurobus 89 — September 4-6th
Novotel Hotel, London. Organised by Microdynamics Inc. UK organisers Pattern Ltd on 01-940 4625.

Circuit Theory And Design — September 5-8th
University of Sussex. Ninth International conference. Contact IEE on 01-240 1871

Holographic Systems, Components And Applications — September 11-13th
University of Bath. Second international conference. Contact IEE on 01-240 1871

POS 89 — September 12-14th
Business Design Centre, Islington, London. Point Of Sale technology exhibition. Contact Batiste Exhibitions on (0532) 580033

Data Communications and Networks — 17-22nd September
Aston University. Vacation school, sixth year. Contact IEE on 01-240 1871 ex 308.

Optical Systems — September 12-14th
Ramada Inn, London.

Sensors And Their Applications — September 25-27th
University of Kent, Canterbury. Conference sponsored by The Institute of Physics, Institute of Measurement and Control, Institution of Mechanical Engineers and IEE. Contact The Institute of Physics on 01-235 6111

Lightning And Static Electicity Conference — September (date to be finalised)
University of Bath. Sponsored by Ministry of Defence Procurement Executive. Contact ERA Technology on (0372) 374496

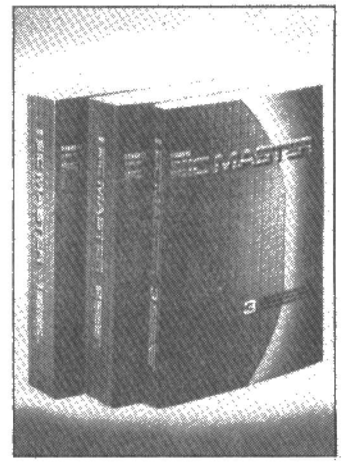
Artificial Neural Networks — October 17-18th
IEE, London. International conference. Contact IEE on 01-240 1871

Open Systems 89 — November 1-3rd
Olympia, London. Computer communications standards. Contact Cahners Exhibtions on 01-948 9800

Automated Test And Diagnosis — April 9-12th
International Centre, Bournemouth. Conference. Contact IEE on 01-240 1871.

BOOKS

1989 IC Master, Vols 1, 2 and 3.
Editor: David Howell.
Published by Hearst Business Communications.
Three volume set £98.00 from J.B. Tratsart Ltd., 154a Greenford Road, Harrow, Middlesex, HA1 3QT.



Weighing in at 4½kg, the massive three-volume IC Master is a brave attempt to catalogue every IC known to man. Well, 80000 of them anyway. If the directory works, it should be possible to find out what almost any IC does, where you can get one, what substitutes or equivalents there are, whether there are any application notes about it (and where to get them from), and a lot more besides.

The amount of data varies widely from IC to IC. What you are guaranteed is a listing according to function in one of the many tables. For the MB3501, for instance, you discover that it's a wideband amplifier made by Fujitsu, but that's your lot. For more information you turn to the distributors directory, find there's no listed UK sales office, and wonder why you didn't spend the £98 on ice cream.

But this is not fair. Most of the time there's enough data available to at least allow a comparison between various similar ICs and make a list of devices that will be suitable for your purposes. The regulators section, for instance, is divided into listings of fixed positive, fixed negative, dual, adjustable positive, and so on. Each listing is further divided according to output voltage and yet again for output current. So if you need an 8.5V regulator good for 1A, you can track it down in no time — there are three to choose from.

Logo Guru

In addition to the main IC listings, there are so many other areas of additional information that it would take a review as long as the IC Master itself to do full justice to them. You'll find a part number guide which not only explains the letters that manufacturers append to the device codes, but also gives the logo or symbol that you might find printed on the IC — a useful first step in identifying an unfamiliar one. Here are a few of the others that caught my attention.

The application notes directory makes fascinating reading. The listed data ranges from the ordinary and predictable (there's scads of stuff on op-amps) to the most arcane, esoteric and outlandish. How about a *Thermometer for Albacore Fishing?* or a *Hands Free Featurephone?* — whatever that may be!

Much less interesting to browse through, but every bit as useful, is the alternate source directory. It's a bit of a letdown to discover that the glamorous sounding XR3503 is nothing more than a workaday LM324 but that's just the thing you'd like to know if you came across it in a piece of equipment you were repairing.

Data

The manufacturers' data section has been greatly extended in this year's edition — most of the second volume is given over to it. It's useful where it appears but since the choice of what data to provide, or indeed whether to supply any at all, appears to be at the discretion of the individual manufacturer, the coverage is very patchy. Some of the smaller concerns list all they've got; other firms (National for instance) give it a miss altogether. If you're lucky you'll find what you're looking for but it's no substitute for a shelf full of data books.

To an individual the £98 price tag will give pause for thought but to anybody who designs circuits, repairs equipment, or is involved in the surplus components trade, I say: go for it! There's an awful lot of information for your money. Large companies, of course, fill their shelves with directories as a matter of course and may even overlook this one because it doesn't absorb enough of the budget. (How can you ask for a bigger budget next year if you haven't overspent a little on this year's?) To you I say: give it a try anyway. You'll find it's the one that actually gets used!

My only real grouse about the latest edition is that the hard cover has been abandoned in favour of a floppy paper one. Making it behave on the bookshelf is about as easy as trying to prop a drunken sailor against a lamp post. But I guess you can't have everything.

Paul Chappell

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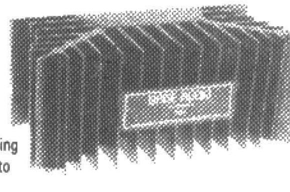

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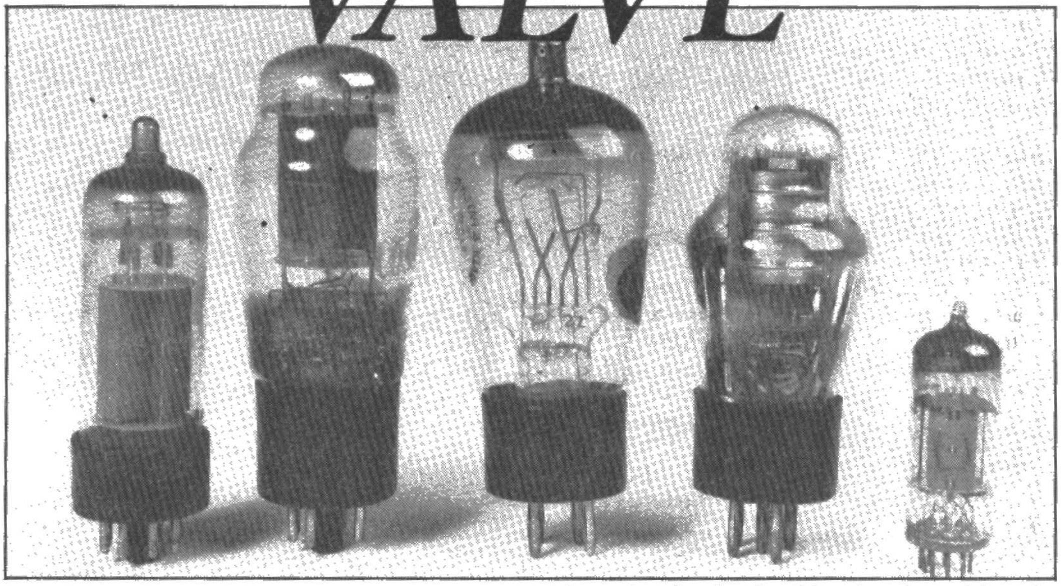
VALVE



John Linsley Hood looks at the history and audio applications of the forsaken but not forgotten thermionic valve

For the first half century of its existence, the art we call 'electronics' was all about the construction and use of electrical circuitry based on 'thermionic valves' — and there are still some remaining enthusiasts, especially in the field of audio, who think that 'valves' are the only proper way of doing things.

There is, quite clearly, an aesthetic appeal in the sight of a row of gleaming bottles with complicated arrangements of wire and metal glowing inside. It is an appeal quite lacking in the dull black plastic blobs and caterpillars of today and one which would encourage a kind of electronic equivalent to the steam railway preservation enthusiasts. But how and why did valves come to be made and used in the first place?

Well, I suppose it all began with James Clerk Maxwell, an Edinburgh mathematician with an interest in physics who had turned his attention to the basic principles of electricity and magnetism. In 1864 he proposed an 'Electromagnetic Theory of Light', the basis of which was that light was just another form of electromagnetic radiation and offered a series of equations which accurately defined the characteristics of all electromagnetic phenomena of this type.

This theory was received with enthusiasm by the scientific community of the day because it tidied up a lot of loose ends which had lain around since the time of Newton. However, like most useful theories it raised as many questions as it answered, one of which was whether all oscillatory electromagnetic fields would indeed cause energy to be radiated through free space, as Maxwell's equations implied.

The Discovery Of Radio Waves

Consideration of this possibility provoked a German professor of physics at Karlsruhe, Heinrich Hertz, to carry out a series of experiments to see whether this effect — the propagation of an electromagnetic wave from some discrete energy source to a remotely positioned receiver — could be demonstrated in the laboratory. In 1889, after some four years of experimentation, he announced that he had conclusively proved that this did happen just as Maxwell had predicted.

The latter years of the nineteenth century became a time of great activity in the field of electrical telegraphy, first by the 'Morse' code and later by speech, and the prospect of being able to send telegraphic messages without the need to install connecting wires — a true 'Wireless Telegraphy' — offered great commercial prospects.

The possibilities were not lost on Marconi and he began his own experiments with 'Hertzian waves' using much improved apparatus. A series of trials commenced, demonstrating the use of such transmissions in places where cables could never be used, such as from ship to shore and from ship to ship, and across the Bristol channel. He mounted his historic 'coup de theatre' by a transmission across the whole of the North Atlantic ocean from Poldhu on the Lizard peninsula in Cornwall to St John's in Newfoundland, which showed beyond doubt the potential for this new discovery.

Understandably, the radio signals received across such distances were very weak and the apparatus used for receiving them was primitive and inefficient. No way was then known of amplifying the signals. So from a strictly commercial point of view, the uncertainties of reception made it a rather dubious competitor to the well-established techniques of telegraphy over wire or undersea cables. For 'wireless' to succeed, it needed improved hardware and this proved a powerful stimulus to invention.

The Thermionic Valve

The thermionic valve really has its origins in the 1880s, when Thomas Edison, the inventor of the filament lamp bulb, had noticed that the glass wall of the bulb would become darkened after prolonged use in regions close to the filament, even though the bulb had been evacuated of all gases.

He also found, as a matter of academic curiosity, that if a metal plate was inserted into the lamp bulb, the glass would stay clear in the area of its shadow. Moreover, if he made this plate positive in respect to the filament then a small current would flow.

This phenomenon was correctly interpreted by Sir Ambrose Fleming as being due to the emission of electrons from the filament and their subsequent capture by the positively charged plate. The significant feature of this effect was that current could only flow in one direction (from the filament to the plate) since

no electrons would be emitted by the cold metal of the plate itself.

This was just what was required to allow the inaudible pulses of high frequency AC received by a Marconi-style 'wireless telegraphy' apparatus to be converted into unidirectional pulses of current which could be heard by headphones or displayed on a sensitive milliammeter. So in 1905 Fleming took out a patent for a lamp bulb with an internal plate, which he called a 'thermionic valve'.

He used this name because 'thermions' was the technical term for thermally emitted electrons and he described it as a 'valve' because of its capacity to allow the current to flow in only one direction.

The next step was taken by an inveterate American experimenter, Lee de Forest. He grasped the fact that since electrons are negatively charged, their flow from the filament to the plate could be controlled if a wire mesh 'grid' was inserted into the space between them. If this was at 0V then the current would flow as before, but if this grid was made sufficiently negative the current flow would be cut off.

This kind of device would for the first time allow small high frequency signals to be amplified and make radio reception a much more reliable business. So in 1907, de Forest took out a patent for this device, which he called the 'Audion' tube, and by so doing launched the whole business of electronics.

The Modern Electronic Valve

Clearly, the thin carbon and tungsten filaments of the early lamp bulbs weren't a good recipe for a copious source of electrons. Tungsten isn't bad but it needs to be very hot (around 2250-2350°C) to be of much use — that consumes a lot of power and doesn't make for a long life. In any case, what was required for valves was that they be heated by a low voltage AC source, derived from the mains by a transformer.

So in contemporary designs, valves have a 'cathode' (the name which is given to the electron emitting electrode) in the form of a metal tube usually made of nickel which is heated by an internal insulated bundle of tungsten wire and is coated with a mixture of barium and strontium oxides.

When this is heated to a dull red heat (about 850°C), a small part of the barium oxide in contact with the nickel is reduced to metallic barium and this diffuses outwards to the cathode surface. However, a very hard vacuum is needed with 'oxide coated' valves to prevent the hot metallic barium from promptly oxidising again!

Metallic barium has a much lower 'work function' (the temperature-related energy which thermally excited electrons have to reach to escape into the vacuum surrounding the cathode) than tungsten and these oxide coated cathodes are now the standard form in valves or 'electron tubes'. I have shown a typical arrangement for an 'indirectly heated' cathode in Fig. 1.

A 'diode' (a two-electrode valve) for use as a detector for AC signals or power rectifier to convert the incoming AC power supply from a transformer into a suitable DC voltage (usually between 100-450V) to operate equipment, is constructed by mounting an electron receiving plate (called the anode) in reasonably close proximity to the cathode.

Usually the anode is in the form of a small diameter tubular sleeve mounted around a cylindrical cathode. It is usually also made of nickel and may be blackened to assist it to radiate heat. To keep the structure rigid, the electrodes will be spot welded on to stiff metal rods, held between the glass 'pinch' at the base and a mica washer which is itself a tight fit in the envelope. With modern tubular-envelope

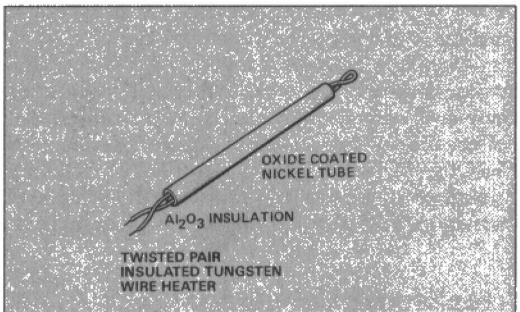


Fig. 1 Cathode structure of an indirectly heated valve

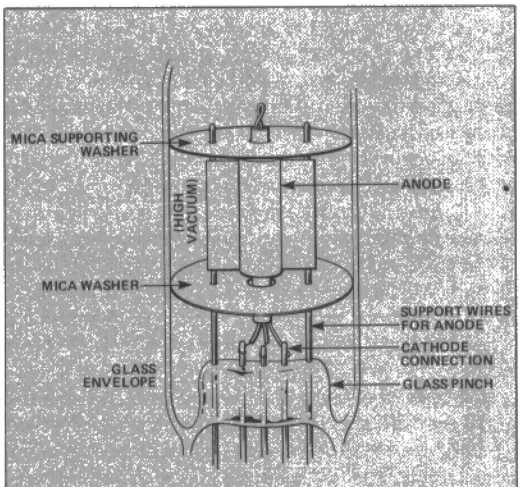


Fig. 2 Mechanical construction of a rectifier diode

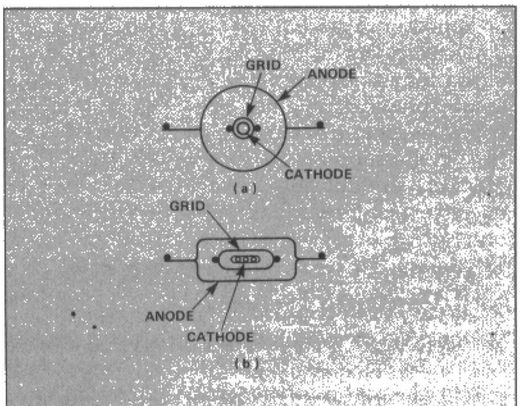


Fig. 3 Comparative structures of triode valves (a) high gain (b) low gain

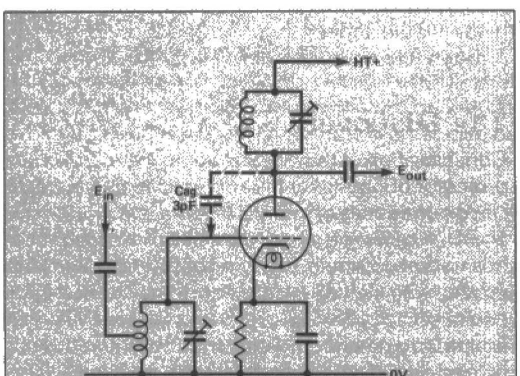


Fig. 4 HF amplifying stage

valves, two such washers are used for even greater electrode rigidity. I have shown this in Fig. 2.

Usually in this type of application, the closer the spacing between the anode and the cathode the better since it will lower the impedance and the consequent voltage drop of the rectifier. Of course the electrodes must not ever touch, even under shock or vibration. Enough of a gap must be left between them in power rectifiers to make sure that they don't spark over inside.

Amplifying Valves

These are all multiple electrode valves with one or more 'grids' or other electrodes between the cathode and the anode. They are called 'triodes', 'tetrodes' or 'pentodes' and so on, according to the number of electrodes (ignoring the heater) which they contain.

Taking the case of the triode, the simplest amplifying valve, the effectiveness of the 'grid' in controlling the current flow through the valve (for a given grid voltage) depends on the fineness of the mesh construction and the relative spacings of the grid and anode with respect to the cathode.

If the grid is relatively close to the cathode and the anode relatively distant, the amplification factor of the valve will be large. Its impedance will however be high and its maximum anode current flow will be small. Obviously the characteristics are reversed for a grid closer to the anode. A comparison of these types is shown in Figs. 3a and 3b.

A couple of illustrative examples of the effect of grid mesh and anode-cathode spacings are given by a pair of double triode designs from the classic 'octal' based series. The 6SL7 has an anode current impedance (Ra) of 44k, an amplification factor (u) of 70, and a normal anode current (Ia) of 2.3mA. Its lower impedance brother, the 6SN7, has an 'Ra' of 7k0, a 'u' of 20, and an 'Ia' of 10.6mA

For power output stages, where the total output power depends on the anode current, the maximum practicable amplification factor of an 'output triode' might only be five or six, so this style of output configuration was seldom used even in 'hi-fi' gear, where the particular warmth of 'triode sound' might have been an incentive.

Screened Grid, Beam Tetrode And Pentode Valves

The main snag with the triode, though, was the capacitance between its anode and its grid, which would be several pF even in a small signal high gain valve. This meant that it was very difficult to arrange HF amplifying stages using triodes, since the kind of circuit I have shown in Fig. 4 would certainly oscillate at the drop of a hat. (It was possible to 'neutralise' this feedback capacitance but that made the circuitry a lot more complicated.)

The answer was to insert another grid, only this time with a suitable positive voltage, between the control grid and the anode to act as an electrostatic screen. This is shown schematically in Fig. 5. This could reduce the anode to control grid capacitance to about 0.005pF and make high orders of RF amplification possible without instability.

This layout also had another advantage in that it allowed a very much higher amplification factor than possible with a triode — figures of the order of several thousands being feasible. However, the snag was that electrons accelerated towards the anode could cause the emission of 'secondary' electrons from the anode, which would be picked up by the positively charged screen grid and cause a 'kink' in the anode current vs anode voltage graph, as I have shown in Fig. 6.

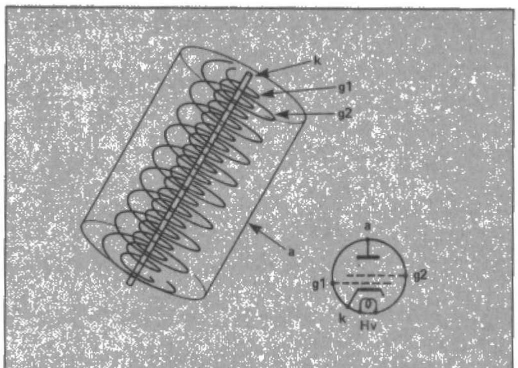


Fig. 5 Structure and circuit symbol for the screened grid valve

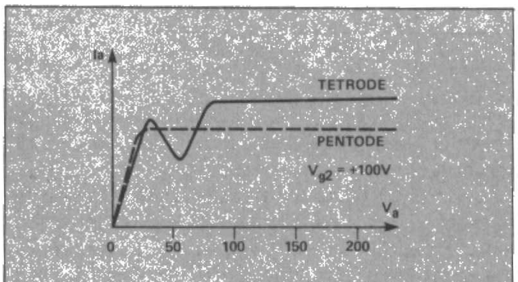


Fig. 6 The characteristic anode current kink of a tetrode valve

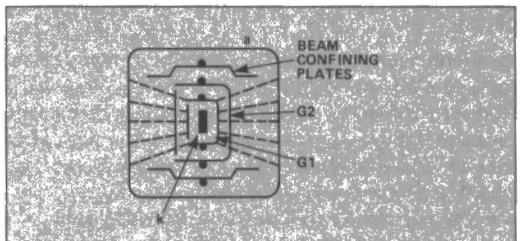


Fig. 7 Beam tetrode structure

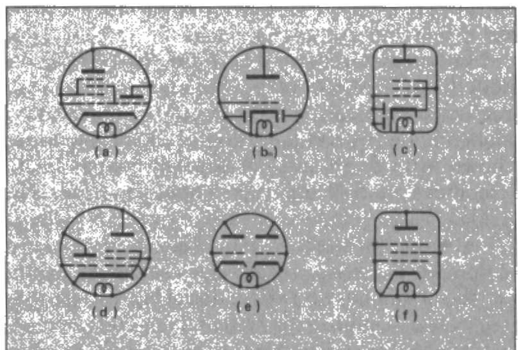


Fig. 8 Circuit symbols for more complex valve structures

This didn't present any great problem for RF amplification where the likely size of the voltage swing present at the anode was unlikely to take the anode voltage into the problem region, but where such valves were to be used for audio (especially in output stages) the presence of the 'tetrode kink' could seriously limit the possible output voltage swing.

Two ways were found of overcoming this difficulty. The first was to interpose a fairly open mesh grid between the screen grid and the anode, connected internally to a low potential (such as the cathode or externally to the zero volt rail). This would produce a reverse voltage gradient close to the anode which would discourage the emission of secondary electrons, while having very little effect on the higher

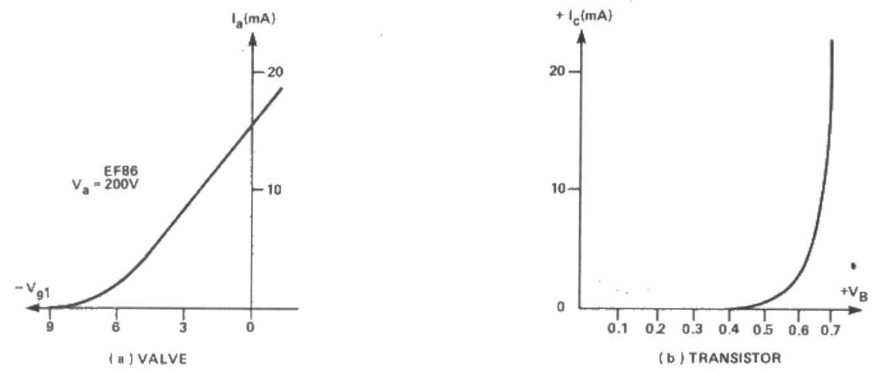


Fig. 9 Input characteristics of (a) valve and (b) transistor

velocity cathode-anode electron flow. This kind of layout was called a 'pentode'.

The second and more elegant answer was to construct a pair of 'beam confining plates' on either side of the cathode-anode electron path as I have shown in Fig. 7. Such valves — called 'beam' or 'kinkless' tetrodes — became the mainstay of the output circuitry of the traditional valve operated audio power amplifiers, since they had high efficiencies and a rather lower 3rd harmonic distortion content than their output pentode predecessors.

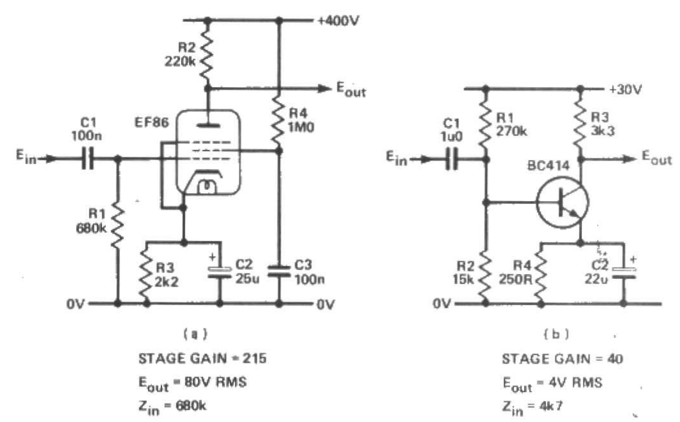


Fig. 10 Comparable valve and transistor gain stages

Other Multiple Electrode Valves

For many years, manufacturers of radio sets had to pay a royalty based on the number of valves that they used in their designs. However even when this royalty payment lapsed, the economics of manufacture favoured multiple electrode valves since fewer valves meant fewer holes in the chassis, fewer valve holders and fewer interconnecting wires. In those days, all interconnections between components had to be made by hand with individual workers sitting, soldering irons in hand, at an assembly bench so that the fewer bits one had to join up the less the job would cost.

The most common multiple valve types were the triode-hexode (shown in Fig. 8a) used for the input 'frequency changer' in a superhet radio, and the double-diode triode and double-diode pentode (shown in Figs. 8b and 8c) used as combined IF demodulator and AF amplifier/output stages for such radios.

In television sets there would also be the triode-pentodes of Fig. 8d for use in time-base circuits, along with the ubiquitous double triodes of Fig. 8e — used very widely in industry, as well as in 'hi-fi' amplifiers. The small signal RF pentode such as the EF80 (shown symbolically in Fig. 8f) was also found to be a very useful device for AF amplification, and a special low-noise low microphony version of this called the EF86 was produced for use in audio equipment.

The last important variation of this device was the 'vari-mu' pentode, sometimes depicted in circuitry by the use of the pentode symbol with an arrow drawn through it. This device allowed the gain of an RF amplifying stage to be controlled by means of a varying negative DC voltage applied to its control grid. This was achieved by winding the control grid with a spiral mesh of wire which became more widely spaced towards one end. Then as the negative 'bias' on the grid was increased, the electron flow through the finely spaced regions of the grid mesh would be cut off and the only remaining electron flow would be through the more open spaced region, which would have a much lower amplification factor.

There were also many specialised combination valves, such as the double anode valves used in high performance mixers and modulators, and the 'magic eye' tubes used as tuning indicators. The main limit on the inventiveness of the valve manufacturers was the eight or nine pins available for such electrode connections on the base of the valve.

Advantages And Snags

The big advantage of the valve is its robustness. It really does take quite a lot of carelessness to damage one physically — except perhaps the sort of user capable of pushing a second compact cassette into the hole of a car cassette player already occupied by the first one (I saw this achieved recently).

Valves are also quite robust in use. They have enough thermal inertia to be able to absorb, for a short time, a current or voltage overload many times their normal rating, and a 1½-2× overload for several hours. They also operate at a high enough anode voltage for quite high output voltage swings to be easily obtained, plus they have a relatively graceful overload characteristic.

In terms of the distortion of the output signal, they are a lot better than any normal semiconductor with the exception of high voltage MOSFETs, and they are a lot better than these in respect of noise level and can't be damaged by 'static'.

On the debit side, there are a lot of problems. To start with, they are physically bulky and soak up a lot of power. Anyone contemplating a twenty valve unit would be envisaging a large hot lump of kit.

Again, the heater wiring has to be taken to each valve holder and carries an AC current at mains frequency and typically at 6.3V. The hum induced by

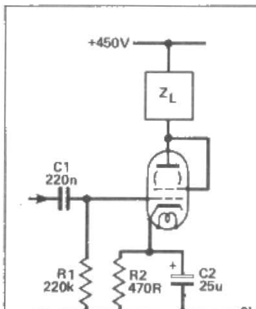


Fig. 11 'Triode connection' of output beam-tetrode

this wiring, especially in small signal circuitry at the input to an amplifier, is a nuisance which must be minimised by care in the layout of these wires, installed as a twisted pair or screened cable (and if the latter, heavy enough to carry 1-2A).

For anyone used to transistor circuitry, the absence of 'complementary' symmetry in the valves is a great pity, since 'upside-down devices can be a great help in neat circuit design.

One also misses the freedom to operate devices at any sort of static voltage level one wishes since, with valves, one must always keep an eye on the limits on cathode-heater insulation and the possibility of electron emission from the heater itself if the cathode is a long way positive of the heater.

The order of voltages which one may encounter in valve operated hardware is also such that one can get a nasty shock if one plays with the innards while the power is switched on, while with semiconductor equipment the damage is more likely to be to the semiconductors.

This leads to the other disadvantage with valves — that their output impedance is so high. This means that it is impracticable to connect a standard 4R or 8R loudspeaker, even a 15R one, directly to an audio amplifier circuit. The necessary output matching transformer, even when carefully and expensively made, is likely to introduce much more degradation of the signal — especially on transients — than the whole of the rest of the circuitry.

The bulk and weight of decently made output transformers, to say nothing of the mains transformer, will mean that a stereo power amplifier is going to be quite a massive bit of gear. Still, people do build them and people do use them. So for the interest of the curious, I propose to continue with a look at the design technology of valve operated audio amplifiers.

Valves In Audio

Nearly thirty years ago, Fairchild Instrument Corporation's invention of the 'silicon planar' transistor process proclaimed to the world that the transistor had come of age and signalled the beginning of the end for the dominance of the thermionic valve in the world of electronics.

Well, as I said, that was nearly thirty years ago. And yet valve operated audio amplifiers are still around and, what is more, they are still being made and sold to audiophiles — so why is this?

As with all things technical, there is good news and bad news. Let us take the good news first.

The Advantages Of Valves

If one is considering the design of low distortion audio amplifiers, the valve has a lot to commend it. To start with, it has a very linear input voltage vs output current transfer characteristic as compared with the silicon transistor, as is shown in Figs. 9a and 9b.

Putting some figures to these graphs, the harmonic distortion at the largest sensible output signal level for a small signal AF valve stage could be of the order of 1-2%. For a transistor gain stage using simple circuitry, this figure would be nearer 10%.

Secondly, the valve amplifier stage will have a very much larger possible output voltage swing, so that for a given AF output the distortion would be proportionately lower anyway.

Thirdly, the problems of 'clipping' in small signal stages are of minor importance. Consider, for example, a normal small-signal AF amplifier pentode, such as the EF86, shown in a working circuit in Fig. 10a. This will operate quite happily at anode voltages

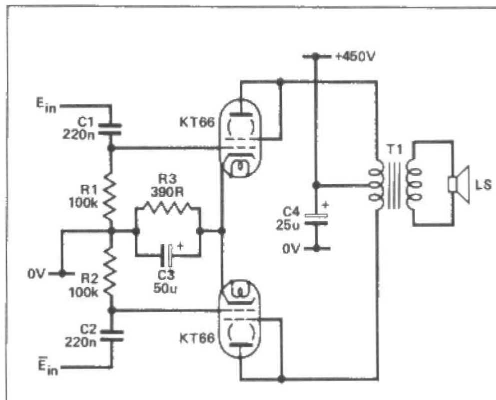


Fig. 12 Typical layout of a high quality push-pull output stage

up to 400V, which would allow a peak-to-peak output voltage swing of some 250V before clipping.

By comparison, a low-noise small-signal transistor, shown in a comparable operating circuit in Fig. 10b, probably has a peak collector voltage rating of some 30-40V and can only deliver an output signal swing one tenth as large before the stage overloads.

'Headroom' is a fairly vital consideration in the preamp stages of solid-state audio gear. One hardly needs to give it a thought with valves.

There is also the question of the stability of performance characteristics. Valves do wear out with time but this is a slow and gradual process and, as a friend of mine once observed, "if you have a valve amplifier and you measure its HT current from time to time, provided that this is of the right sort of order, you know your gear is all right. You can never be so sure with these new-fangled transistor systems".

Finally, provided that one doesn't hit them with something heavy and hard, or take an excessive anode current for too long at any one time, the valve is almost indestructible.

These valve properties were a very great advantage for audio equipment, since the peak to mean signal (and power) ratios of typical music signals are very high and use at a quite modest average output may lead to occasional quite brief peaks in signal level which could cause a transistor amplifier some trouble — especially if the load impedance is a bit on the low side — but which valve gear will shrug off.

This aspect is part of what one might describe as the valve's more graceful overload characteristics, which endears this technology to its afficionados.

The Snags

I have been around long enough in the field of electronic circuit design to be able to remember quite clearly the advent of the transistor, and the impact of its intrusion into my orderly hot-cathode scene. Obviously there were problems in the use of these little gadgets — normal valve type circuitry didn't work terribly well with them.

Indeed, most of my own early trials with transistors seemed to end up with a very sick semiconductor, from which a small ascending spiral of smoke signalled the end of the experiment. However, once one did get the hang of their needs and prohibitions, these little gadgets did make clear some of the inconveniences in the use of the valves with which we had lived so long.

To begin with, valves were BIG. Making small sized bits of gadgetry with them was difficult. They also didn't like knocks or vibration. This wasn't much of a problem for domestic audio, with one's hi-fi amp sitting quietly on the book case, but at the time I was an

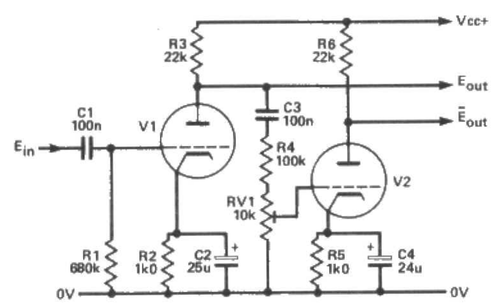


Fig. 13 Simple cascade phase inverter

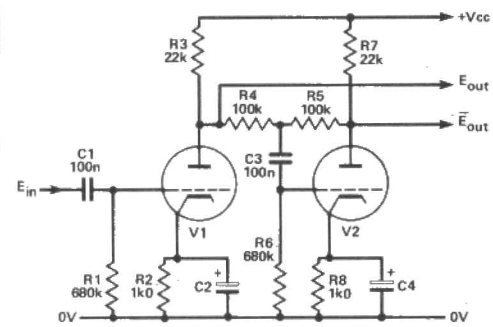


Fig. 14 'Floating paraphase' inverter stage

industrial electronics engineer and the need to make things which would work in hostile environments was all part of the scene.

Thirdly, valves were a bit awkward in use. To make any bit of kit using valves, one first had to start by putting valve holders in a metal chassis, for which one had to begin by cutting the holes. Then one had to wire up the heater circuits using twisted pair wires to minimise the extent of the radiated 50Hz AC field, and then connect them to the mains transformer — for which one had to cut a rectangular hole in the chassis.

With the orders of voltage one needed to use with valves, it was not considered good practice to have the transformer mounted on top of the chassis with its high voltage connections exposed, ready to bite the unwary hand.

Finally, however careful one was with heater circuit wiring or HT smoothing, one always seemed to get a bit of residual mains 'hum' in the output. The ideal result — so easy to achieve with transistors — of an audio amplifier where one needed to look at the pilot light to see if it was 'on', was a difficult thing to achieve.

Nevertheless, a good valve operated audio amplifier was a nice thing to listen to, especially in the fond remembrance of nostalgia, a few decades in the past. I said a 'good' one, since there were some pretty grotty designs, even among those sold with a hi-fi label and now considered to be collectors pieces.

Valve Audio Circuit Design

In any audio power amplifier, the task is to get as much output as one can for a given rating and cost of power supply or output valves, with as low a distortion figure as is practicable.

For small signal amplification, triodes are better than pentodes in that their distortion content is lower and consists mostly of second harmonic, which the ear does not object to (indeed there is some evidence that half a percent or so of second harmonic distortion may actually appeal to the listener. Third harmonic distortion is a major component of the pentode output distortion figure and is definitely not liked by the ear).

Unfortunately, while triodes make quite good (though lowish gain) small power amplifiers, they make very inefficient (though nice sounding) output stages.

The 'beam tetrode' is a style of valve in which the unwanted secondary emission of electrons from the anode (which can cause a big 'kink' in the anode voltage/anode current graph when the anode voltage swings below that on the screen grid) is prevented by a pair of internal 'beam' plates. It is at least as efficient as the pentode and is somewhere between the pentode and the triode in its distortion characteristics.

Power output triodes, in addition to being inefficient, were always awkward to use and are now no longer made, since one can get substantially the same result by connecting the anode to the screen grid (G2) in a pentode or beam tetrode, which avoids the need for the manufacture of a separate valve type.

Using a single 'triode-connected' output beam tetrode as I have shown in Fig. 11 will only give an output power of 4-5W, even with an HT supply of 450V. This isn't adequate for 'hi-fi' use.

The 'push-pull' connection of a pair of triode-connected beam tetrodes such as the celebrated KT66s, as I have shown in Fig. 12, will allow an output power of about 15W RMS, provided that the output stage is matched to the load.

This brings me to the final snag for the audio designer, which is that because valves have a high output impedance (typically of the order of k-ohms), an output coupling transformer T1 is needed to match such an amplifier design to a loudspeaker load. On the quality of this component hangs the whole performance of the design.

Push-pull Output Stages

Using the output valves in 'push-pull' gives a useful increase in output power from this stage and will cancel out most of the even harmonic distortion residues. To do this it is necessary to provide an antiphase pair of input drive signals to the two control grids, so that as one goes positive the other will swing negative.

The simplest way of doing this is by means of an input 'driver' transformer but the use of two coupling transformers in cascade is not a very 'hi-fi' approach. Respectable designs must therefore contrive some kind of 'phase-splitter' circuit, to convert the 'single ended' input signal into a pair of balanced low-distortion push-pull signals, to drive the output valve pair.

It is in the design of the phase-splitter stage and the succeeding amplifier stages which will drive the output valves that the ingenuity of the designer is most needed, since it is a difficult part of the circuit to do well.

I have shown some of the circuit options in Figs. 13-16. Of these the first and crudest, shown in Fig. 13, is just an additional (phase-inverting) amplifier stage V2, with a trimmer pot on its input so that its output can be adjusted to be the same as that from V1. This works but offers few concessions to hi-fi.

An improved version of this circuit, shown in Fig. 14, is called the 'floating paraphase' and uses a pair of resistors (R4/R5) connected between the

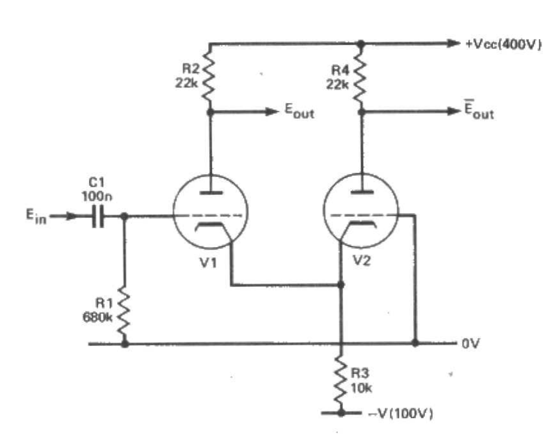


Fig. 15 'Long tailed pair' inverter stage

anodes of V1 and V2 to generate an input signal to V2. Because of the negative feedback through R5, the V2 output will be close to that from V1. Also, because of the nearly 100% negative feedback applied across V2 through R5, the output from V2 will not be much degraded by comparison with that from V1.

The third option in Fig. 15 is our old friend the 'long-tailed pair', very commonly used in transistor and IC circuit layouts. This works best if there is a suitable low-current negative supply to feed the cathode circuit, so that both of the input grids can operate at 0V line potential.

This type of phase-splitter has low distortion but may not, for a single long-tailed pair stage, give identical but antiphase outputs from both V1 and V2. Running two successive long-tailed pair stages in cascade will ensure that the two push-pull outputs are closely identical.

The final option of Fig. 16 is the split-load phase-splitter. In this the second valve V2 is operated as a cathode follower, but with an equal value of load resistor in both its cathode and anode circuits. Since the same current flows through both, the outputs must be identical — apart from the minor effects of the different load capacitances on the anode and cathode circuits.

In practical audio power amplifier designs, combinations of these circuits will be used but of the choices the last option is very much the purists approach. This was the phase-splitter layout employed by D T N Williamson of the Marconi-Osram Valve Co in 1947, in his justly celebrated 'Williamson' 15W amplifier, whose performance has hardly been bettered in valve operated amplifiers to this day.

Practical Problems

The influence of the heater circuit on the operation of a valve cannot be entirely ignored, since there will inevitably be some current leakage between the heater winding and the cathode tube which surrounds it. Also, since there will be some capacitance between heater and cathode, some mains 'hum' voltage will be coupled into the cathode circuit from the heater winding.

Finally, if the cathode is significantly positive with respect to the heater, there may be some thermionic emission from the heater to the cathode and this will inject an unwanted modulated DC into this circuit. This problem could be even worse if the heater was significantly positive in respect to the cathode, when it would operate as another 'anode'.

This difficulty was avoided in traditional designs by the use of a separate heater supply winding on the mains transformer, which could be connected to a suitable positive voltage supply so that the heater element was that the same potential as the cathode.

Output Stage Efficiency

The output efficiency of push-pull triode-connected beam tetrodes is nowhere near as good as that of the same valves when tetrode connected, and this had stimulated some research during the early 1950s into ways by which the output stage efficiency could be improved (without too great a worsening of the THD figure) by coupling the cathode or screen grid circuit into the output transformer.

Various ways of doing this were tried — using additional windings or taps on the windings for this transformer — but the simplest and generally most effective technique was found to be to connect the screen grids (G2) to a tap on the primary winding, somewhere between 20% and 40% of the total (Fig. 17). This gave an output stage distortion which was nearly as good as that of the same valve when triode connected but without causing too big a loss in efficiency.

This technique was described as the 'ultra-linear' connection. I remember an indignant purist of the day feeling that nothing could be more linear than 'linear', observing of this description that "it was like the thirteenth stroke of a crazy clock . . . which cast doubt on all that which had gone before". Nevertheless, this was a useful improvement in output stage circuitry and allowed existing valves to be used in 25-30W amplifiers.

Negative Feedback

Most of the exponents of valve operated audio amplifier technology are also committed to the view that negative feedback is a 'bad thing' and must be avoided. However, as a concession to the main stream of circuit design thought, they may allow some 6dBs worth in their designs. By doing this they fall between two stools.

The purpose of negative feedback is threefold. Firstly, it improves the accuracy of waveform preservation and this will apply, within certain bandwidth constraints, to transient signals as well as steady state 'sine wave' ones.

Secondly, negative feedback will help stabilise the gain, from unit to production unit and from LF to HF. Thirdly, it will reduce the amount of unwanted 'hum' and noise introduced by that part of the circuit within the feedback loop. However, to do any of these things properly it must be adequate in extent. This is particularly true for the reduction in waveform distortion, for the following reason.

Assume that the amplifier introduces a certain amount of second and third harmonic distortion. Together with the original signal, this will be fed back and added to the input but in antiphase. Because the distortion components are not present in the input signal, they will be amplified again to cancel the original errors.

However, and this is the important bit, the distortion components fed into the circuit by the feedback loop will also be distorted, so the original second and third harmonics will now generate fourth, sixth and ninth harmonics. These last two are aurally objectionable.

If enough feedback is used, the residual magnitude of the internally generated high order harmonic distortion will be vanishingly small and won't go round the loop for a third and fourth time creating yet more spurious overtones. However, to be on the safe side at least 20dB, preferably 26dB, of NFB must be used

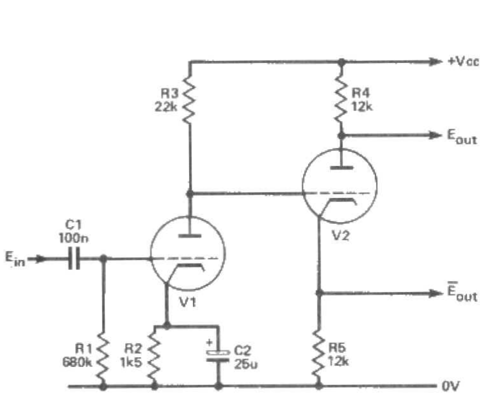


Fig. 16 Split-load phase splitter circuit

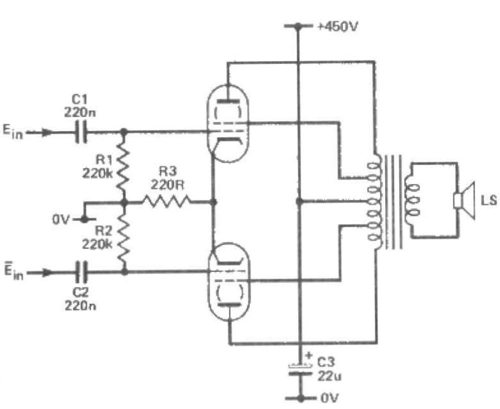


Fig. 17 The 'ultra-linear' output stage configuration

and this demands a very good output transformer — since this is the component which introduces the bulk of the end of passband phase shifts.

Power Supplies

A high power audio amplifier will normally be operated from a single +450V HT supply, and this must be adequately smoothed to keep the hum level down. Fortunately for those who make modern day valve operated audio amps, reasonably sized electrolytics are now much more easily obtained than in the 1950s and this helps a lot.

The smoothing effect of a capacitor depends on the energy which is stored in it, given by $CV^2/2$. This means an 8µF capacitor at 450V holds about as much energy as a 1500µF at 35V. If one were to 'push the boat out' and use a 100µF/450V smoothing capacitor, this would be equivalent to a 16,500µF at 35V. Even the advocates of filing cabinet sized reservoir capacitors would probably think that this was big enough.

However, a 450V/100µF capacitor would pack a lethal punch, and the HT+ supply line would need to be treated with great respect. In addition, an inadvertent short-circuit across the HT+ line would be noisy and highly destructive.

A Present Day 30W Valve Audio Amplifier Design

I have sketched out in Fig. 18 the type of circuit which might have been used when valve amplifiers were in their heyday if some of the components we now take for granted had been available to their designers.

This is largely based on the Williamson design but with an 'ultra-linear' connected output transformer and fixed bias for the output valves, derived from a three-terminal voltage regulator in the interests of output stage efficiency.

The performance of the design will depend, crucially, upon the characteristics of the output transformer. Assuming that this has a specification which is as good as that of the original Williamson one, the indicated amount of negative feedback will be usable to give a bandwidth of 3Hz-100kHz, with a full output THD of 0.1% at 100Hz-20kHz, decreasing at lower output powers to below the noise threshold at 1W or so.

I have only considered power amp designs because, while it is possible to make valve operated pre-amplifiers as well, I really cannot see any good reason for such a masochistic exercise — except possibly for 'headroom', and this could be achieved more easily in other ways.

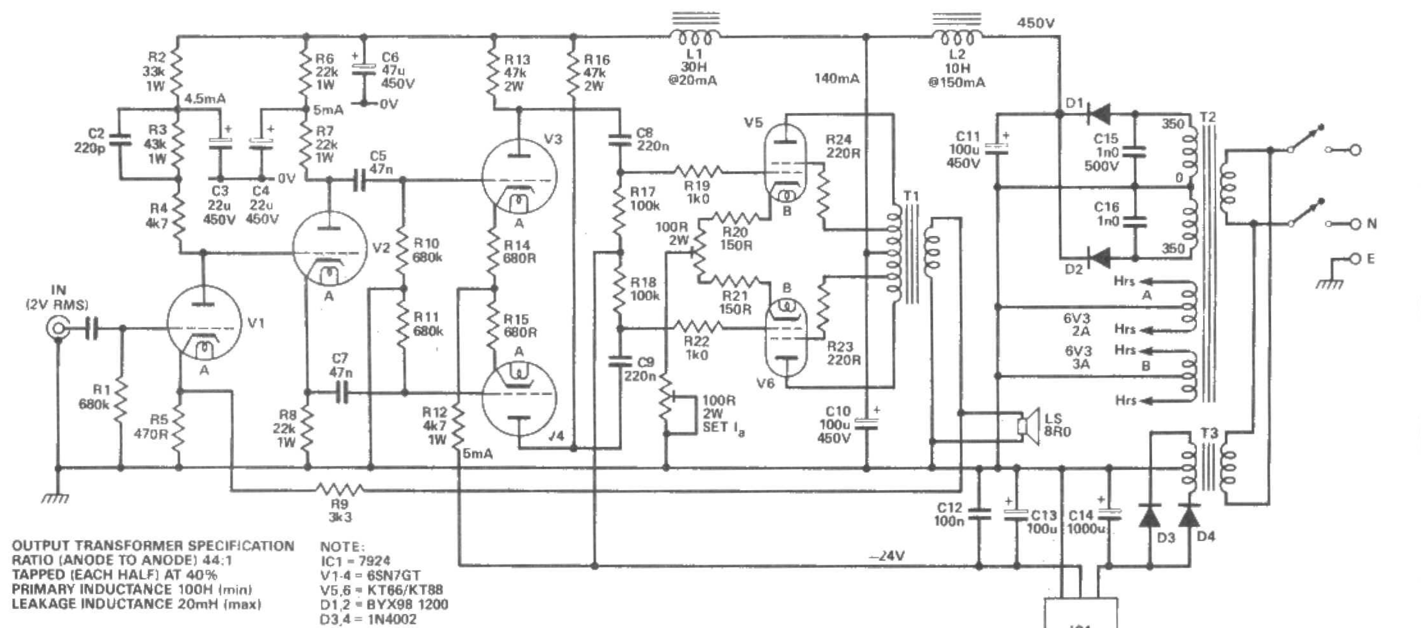
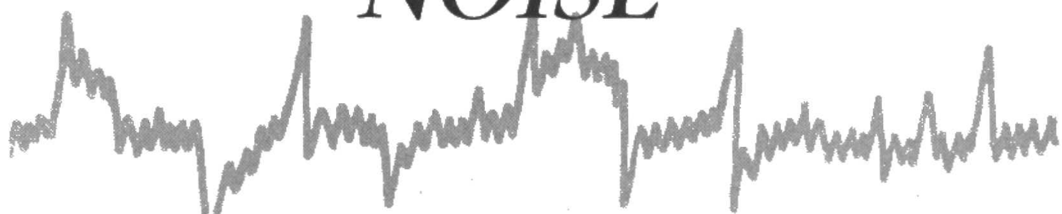


Fig. 18 A high quality 25-30W audio amplifier

NOISE ANNOYS



Paul Chappell mounts a special investigation into the misleading noise characteristics quoted to unsuspecting punters

If you choose your op-amps from catalogue information, or even from some manufacturers' data books, about the only indication you'll get of the noise performance is a figure of so many nV per $\sqrt{Hz}$. Thumbing through the Electromail catalogue, I see that the NE5534 turns in a figure of 3.5nV per $\sqrt{Hz}$, the TL071 gives 18nV per $\sqrt{Hz}$ and the OP27G has a figure of 3.2nV per $\sqrt{Hz}$. So for a low noise amplifier the NE5534 is much better than the TL071, but just a touch inferior to the OP27G, right? Or is there more to it than that?

In the very same catalogue I find that the ZN549CP comes up with a white noise voltage figure of a tiny 800pV per $\sqrt{Hz}$, but that its LF spot noise is 3nV per $\sqrt{Hz}$. Both figures subject to a zero source resistance. What's going on here?

If they had just put down the 800pV figure, my choice would have been clear: 800pV is less than 3.2nV. So I choose the IC with the nicest number for noise and go for the ZN549. But they confuse the issue by putting in another noise number and then say that the whole lot might not be true anyway: it only works for a zero source resistance! Knowing that IC manufacturers are just as keen as anyone else to say their product washes whiter than white, you can bet that the zero source resistance turns in the best achievable figure. But what if the source resistance isn't zero? Will it make a lot of difference? And how about the other ICs — could it be that their figures depend on having zero source resistance and the manufacturers just forgot to mention it? It's enough to make you read the rest of the article!

Voltage Noise Density

Let's go back to resistors for a moment. Take a 100k resistor ar room temperature. By the formula for the noise voltage v_n(RMS) = $\sqrt{4kTRB}$, it's easy enough to calculate that the voltage noise in a given bandwidth will be 40nV × $\sqrt{B}$; that is 40nV times the square root of the number of Hz in the bandwidth, or even more simply: 40nV per $\sqrt{Hz}$.

This means that for a 1Hz bandwidth at any centre frequency whatsoever, the RMS noise voltage from a 100k resistor will be 40nV. The 40nV per $\sqrt{Hz}$ is the noise density figure for the resistor: it tells you how much noise in a unit bandwidth. Since it is independent of frequency, the one figure tells you all you need to know. It can be summed up in the graph of voltage noise density against frequency shown in Fig. 1.

As a slight digression, the 40nV figure for a 100k resistor (or 4nV for a 1k resistor) is a good one to commit to memory. As long as you remember that the voltage noise is proportional to the *square root* of the resistance (so four times the resistance will produce twice the noise voltage, one hundred times the resistance will give ten times the noise voltage and so on) it's fairly easy to do a quick calculation of the noise density figure for any resistor in your head.

Now, what are we to make of similar looking figures quoted for op-amps? The first thing to realise is that the quoted figures are referred to the input of the op-amp. If you make up an amplifier with a gain of say twenty, and make sure that the input sees exactly the same conditions it did when the tests which produced the figures were made, and if you're lucky enough to buy a device which lives up to the 'typical' figures, the best you can hope for is a voltage noise density of twenty times the published figure at the output.

The second assumption you have to make, if you want to rely on the published figures, is that the op-amp has the same noise spectrum as a resistor: equal power per unit bandwidth or 'white noise'. If not, the graph of noise density against frequency will not be a horizontal line, the noise density figure will vary with frequency and the published number won't be telling

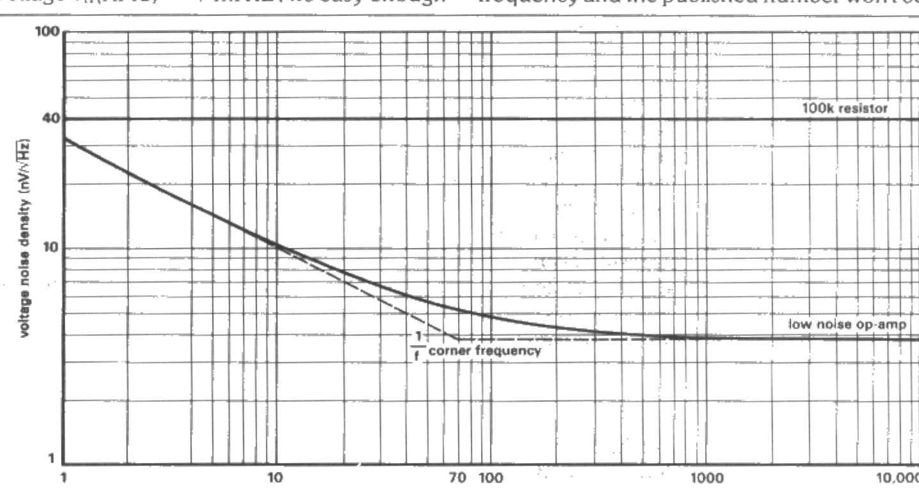


Fig. 1 Graph of voltage noise density against frequency for a resistor and low noise op-amp

CIRCUIT THEORY

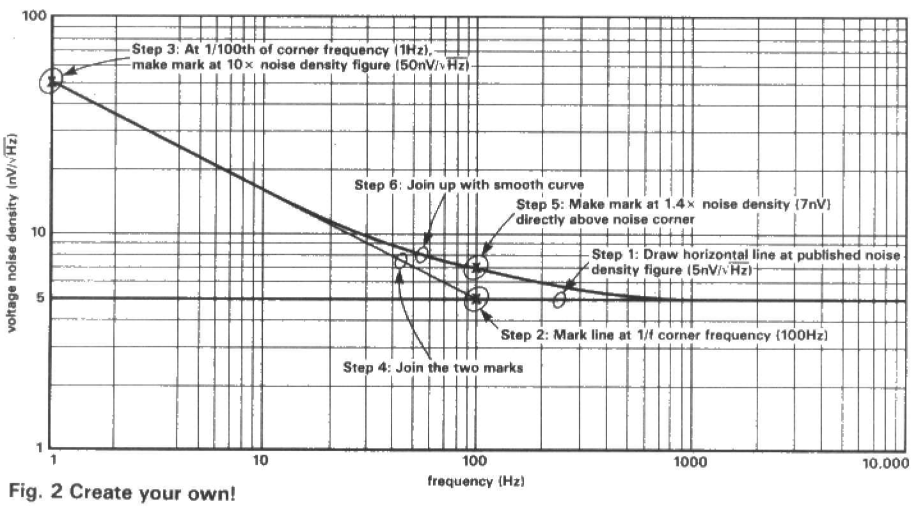


Fig. 2 Create your own!

the whole story! I've already given enough hints in past articles for you to guess that the noise spectrum is not flat for op-amps. The true picture for a low noise audio amp (NE5534s and the like) is also shown in Fig. 1. For 1kHz and above, the 3.5nV per √Hz is a reasonable approximation to the truth. But it's not the whole truth.

The rise at the lower end of the spectrum is caused by 1/f noise, so called because its density is inversely proportional to frequency. At the upper end, white noise predominates. The changeover point (where the two asymptotes join, as shown by the dotted line in Fig. 1) is called the *noise corner* or *1/f corner* of the op-amp. Above this frequency you can believe the published noise voltage figure. Below it, don't trust it an inch!

If you can get hold of the figure for the 1/f corner of the op-amp you're considering, you can rough out a diagram like Fig. 1 for yourself. First, draw a horizontal line at the published noise voltage figure. Let's say it's 5nV per √Hz: you draw a horizontal line at 5nV. Now mark the line at the corner frequency. Let's say it's 100Hz. Figure 2a shows these steps.

At a frequency of one hundredth of the corner frequency (1Hz in this case), make a mark at ten times the published noise voltage figure (50nV) and make a straight line between this and the corner frequency mark on the horizontal line (Fig. 2b). Now make a mark vertically above the place where the two lines join, at about 1½ times the published noise voltage figure (7nV will be fine).

Now you draw in the curve freehand! Start along the downward sloping line until you reach a tenth of the corner frequency (10Hz in this case). Then move away from the line so that your graph will go through the '1½-times' point (or 1.4 times, to be a little more accurate). Then head back towards the horizontal line and join in at about ten times the corner frequency (1kHz). Follow along the horizontal line to the end of the graph and you're done. Given the uncertainty about 'typical' figures and variations in corner frequency between one IC and another, your home-made performance graph will give results every bit as good as one you'll find in a data book! Figure 2c shows the result. Remember that you have to do this on log-log graph paper or it won't give the proper answers.

Current Noise

Even when you've taken account of the 1/f noise, this is by no means the whole story. Superimposed on the op-amp's bias currents is current noise: random fluctuations of the current at the input terminals. You'll usually find a figure for this in data books but not always in catalogues. It will be given as so many pA per √Hz. For the OP37 for instance, it is 0.4pA per √Hz — quite a respectable figure for a reasonably priced op-amp.

The same warning about low frequencies applies here too: there will be a 1/f corner above which white noise will predominate and the noise density figure will be fairly reliable, but below which it will rise. The situation is worse than for voltage noise because the corner frequency is usually very much higher.

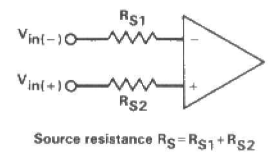


Fig. 3 The bare bones of an op-amp input stage

Figure 3 shows the barest bones of the input circuit of an op-amp. In series with each input is a resistor (which might be zero resistance if the terminal is directly connected to say 0V). The resistors might be physical devices, or could be the source resistance of whatever is providing the input signal, or a combination of the two.

The three things to be taken into account are the op-amp's voltage noise, its current noise and the noise generated by the resistors themselves. The effect of the current noise will be to produce a voltage across the resistors, so the whole lot can be summed up as:

$$\text{total noise} = \sqrt{(\text{voltage noise})^2 + (\text{current noise})^2 + (\text{resistor noise})^2}$$

You'll notice that if you assume $R_S = 0$, you cut out all the contribution from current noise and from resistor noise and get a very cheerful looking figure. It won't bear much relation to the op-amp's performance in a real circuit but it might persuade a few dummies to buy one!

One decision an op-amp designer has to make is how to trade off current noise against voltage noise. An op-amp intended to work with high source resistances will have the current noise minimised, probably at the expense of the voltage noise but since current noise will predominate over a certain value of R_S, this is exactly what is required.

Let's see how this works out. First of all, notice that the 'root of sum of squares' addition process for noise sources means that the larger sources have a disproportionately big effect on the total. Let's suppose that there are two noise sources, one of 3nV and one of 1nV. The total noise will be $\sqrt{9+1}$ = 3.16mV, so neglecting the 1nV entirely would only give a little over 5% error in your estimate of the total noise! Unless the two sources are very similar in size, there's little error in saying that all the noise comes from the larger.

Now, suppose the two noise sources are the voltage noise of the op-amp and the current noise

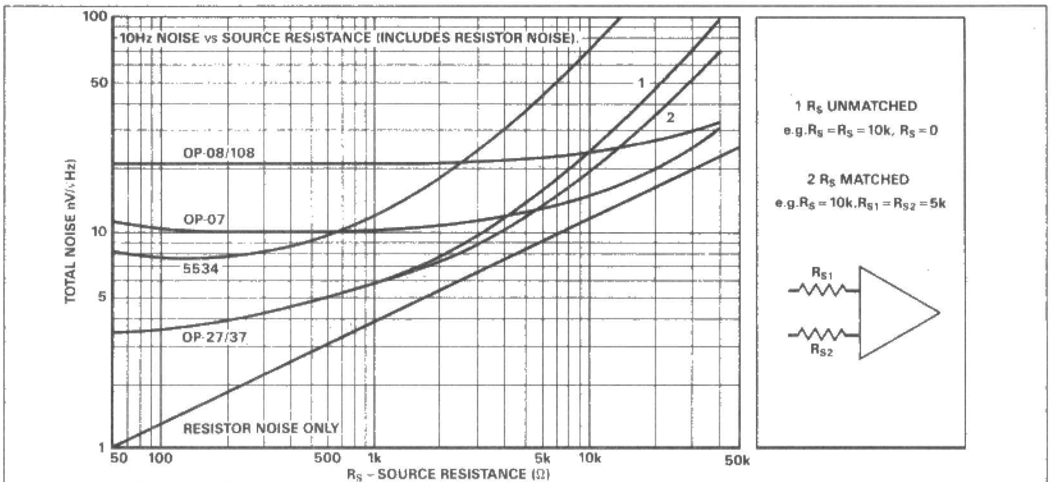


Fig. 4 Noise at 10Hz versus source resistance for various op-amps. Spot the odd one out!

multiplied by the source resistance. If the source resistance is low, the voltage noise will predominate. The current noise will only start having an impact when R_S is large enough for R_S times the current noise to be a third or more of the voltage noise. Taking the figures for the OP37, the voltage noise will be 3nV per √Hz, the extra voltage noise introduced by the current noise flowing through the source resistance will be $(0.4 \times R_s)$pV per √Hz (because the noise current density is 0.4pA per √Hz), so for a source resistance below about 2.5k (which would give 1nV per √Hz) the voltage noise swamps the current noise. Above about 20k the voltage noise is lost in the current noise. For high source resistances then, it's wise to accept a high noise voltage figure for the op-amp as long as it turns in a good noise performance, because it's the current that will make the difference.

To get the total noise, you have to add in the resistor noise, of course. I'm assuming that the source resistance is high and the designer had no choice in the matter (because the input comes from a transducer for instance). If there is a possibility of varying the source resistance, the very best results will come from the lower source resistance, lowest noise voltage, and a more modest figure for the noise current.

The sum of all the various noise voltages for several different op-amps is shown in Fig.4 (taken from the PMI Linear data book). Bearing in mind that it's for a frequency of 10Hz and that most of the PMI op-amps featured are intended for instrumentation and have exceptionally low 1/f corner frequencies, would I be justified in using this diagram alone to 'prove' that the OP27 and OP37 are better, noise-wise, than the Signetics NE5534? PMI don't say this, by the way, but they do slip the Signetics IC in amongst their own data and leave you to draw your own conclusions!

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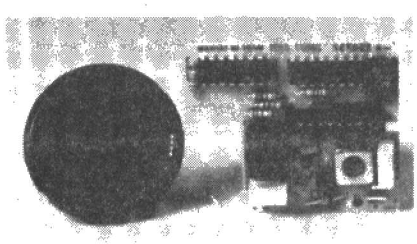
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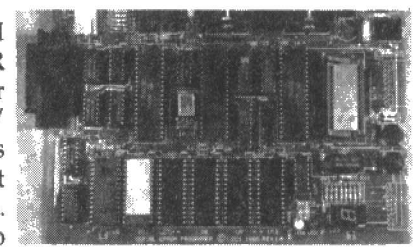

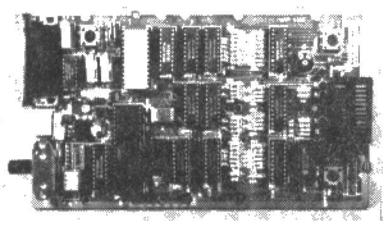
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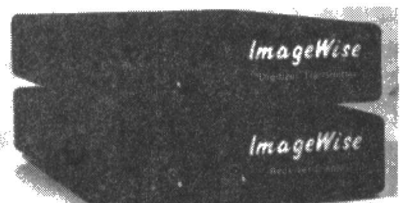

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DC MOTOR CONTROL

Ray Marston opens his circuit notebook to ETI readers, starting a mini-series on DC motors and how to keep control

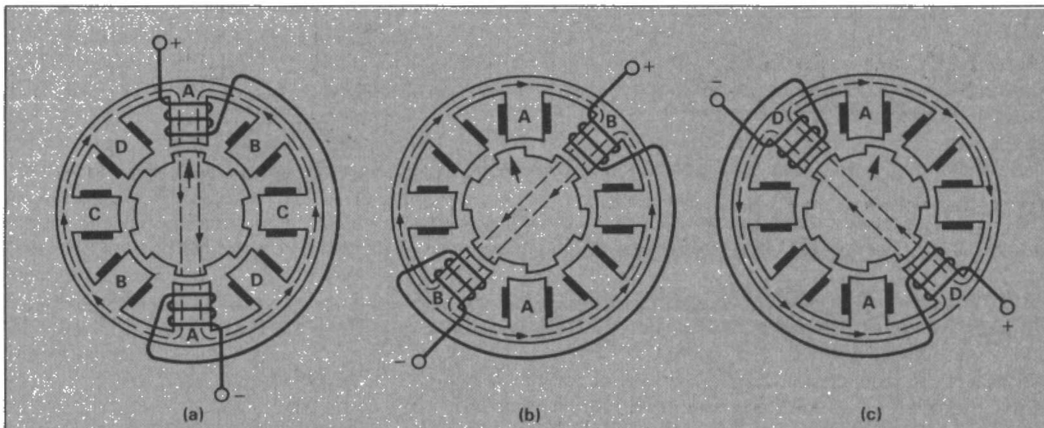

Fig. 1 Motor action of a four phase variable-reluctance stepper motor

One of the most interesting applications of electronics is the control of DC electric motors. Careful techniques can be used to give precision step rotation or speed/direction control of multi-phase stepper motors, or precision speed regulation or wide-range speed control of permanent magnet DC motors, or precision control of the speed or angular movement of various types of DC servomotor, and so on. Practical application circuits of all these types are shown in this and the next part of this mini-series. Let's start off by looking at the various types of DC motor.

DC Motor Types

Four major types of DC electric motor are relevant to us. The first of these is the 'stepper' motor, which has several phase windings. These are pulsed in an appropriate sequence to rotate the output spindle a precise step angle (usually between 1.8° and 7.5°). Thus the spindle can be turned an exact number of steps or rotated continuously at any desired speed in either direction, simply by applying suitable pulse sequences.

Stepper motors can easily be controlled via a microprocessor or dedicated stepper motor driver IC such as the SAA1027 or SAA1024. They are widely used in applications where precise amounts of angular movement are needed, such as in the movement of robot arms, in daisy wheel character selection, or the movement control of the print head and paper feed in an electronic typewriter.

The most widely used type of DC motor is the permanent magnet commutator type, simply designed to rotate at some approximate speed when powered by a particular DC voltage. Motors of this type are used as fixed-speed drivers in tape/cassette recorders and record/disc players, and as wide-range variable-speed drivers in miniature electric drills and model locomotives. In all of these applications, the motor performance can be greatly enhanced with the aid of electronic control circuitry.

The third type of motor is the so-called 'servo-motor'. This is simply an electric motor coupled to a movement-to-data translator such as a shaft-mounted tachogenerator (which gives an output proportional to the motor speed) or a gearbox-driven potentiometer (which gives an output proportional to the output shaft position). When one of these motors is coupled into a suitable power control feedback loop, its speed can be precisely locked to that of an external frequency generator, or its shaft movements can be locked to that of an external shaft or control knob.

Servomotors of the tachogenerator type are often used to give precision speed control of record/disc turntables. Servomotors of the 'pot output' type are widely used to give remote-controlled antenna rotation or remote activation of model aircraft/boat control surfaces and engine speed.

The final type of motor is the 2-phase low-voltage AC motor, usually driven via a DC-powered low frequency oscillator. Motors of this type are occasionally used in turntable driving applications.

Stepper Motor Basics

Stepper motors come in two basic forms, either 'variable-reluctance' or 'hybrid' types. Stepper motor basic principles can best be understood by looking at Fig. 1, which shows stepping operation of a 4-phase variable-reluctance motor.

The stator (body) of this motor has eight inward-projecting teeth, each with a coil winding connected to oppose the coil on the opposite tooth. This gives four-phase pairs: A,B,C and D.

When a phase is energised, magnetic flux flows from the positive phase tooth to the negative one via the shortest possible magnetic path through the soft-iron rotor, which has six (in this case) projecting teeth. To minimise this magnetic path the rotor is forced to move so that the nearest pair of its teeth align with those of the energised phase.

Thus in Fig. 1a, phase A is energised and the rotor's reference tooth (indicated by the large arrow) aligns with positive phase tooth A. From this position

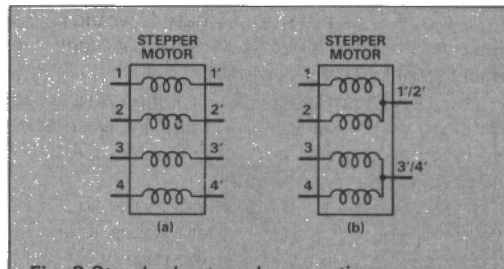

Fig. 2 Standard external connections for a four phase stepper motor

CIRCUITS

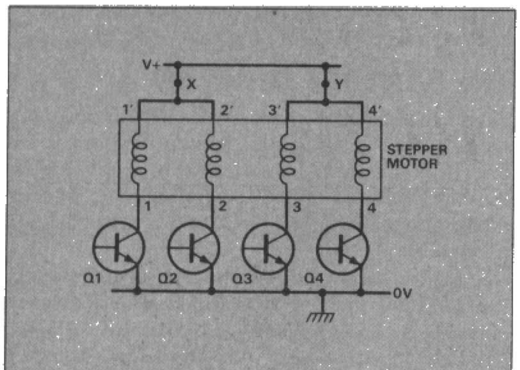

Fig. 3 Basic transistor driven stepper motor circuit

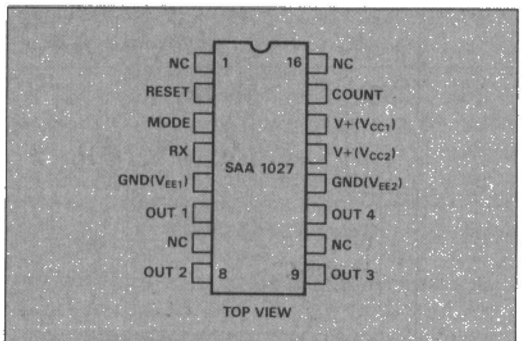

Fig. 4 Outline and pin designations of the SAA1027 stepper driver IC

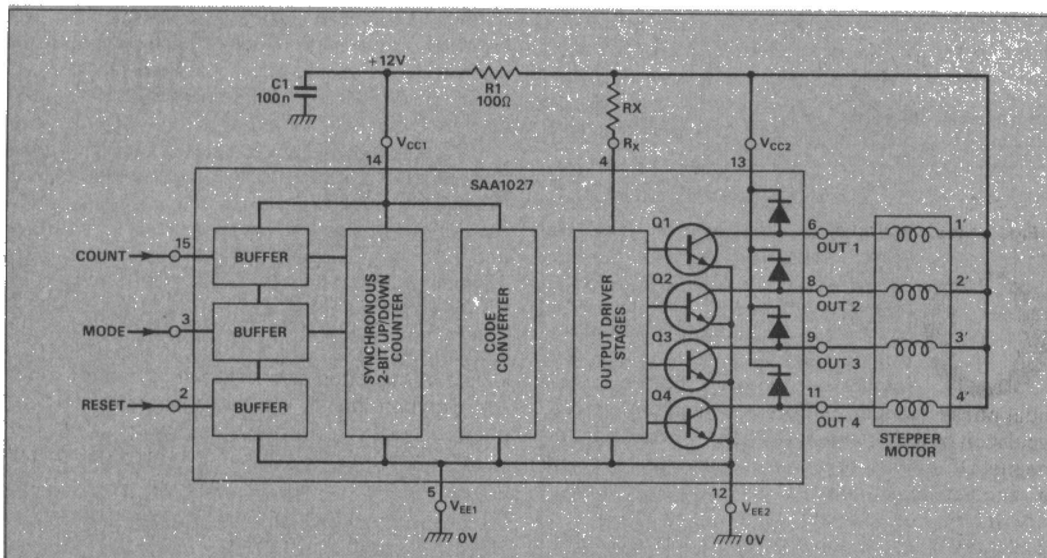

Fig. 5 Internal block diagram and basic applications circuit of the SAA1027

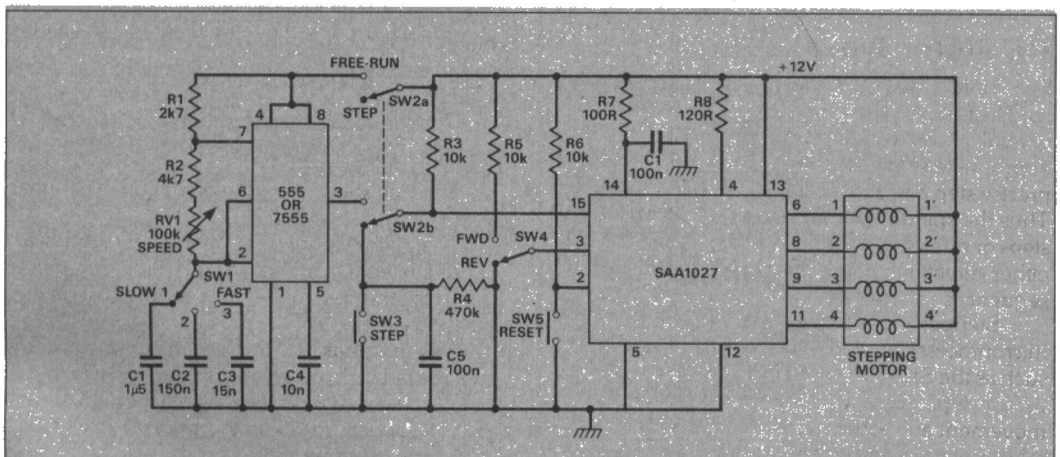

Fig. 6 Complete stepper motor drive and test circuit

we could energise phase B (the rotor turns 15° anticlockwise for the nearest tooth to line up — Fig. 1b) or phase D (rotor turns 15° clockwise — Fig. 1c).

Going from phase A to phase C would give an uncertain result since two possible teeth are equidistant from the C phase.

So each step gives a 15° rotation and we can choose between clockwise and anticlockwise simply by energising in the order A D C B A or A B C D respectively.

The step length of this type of motor equals $\frac{360°}{P \times N}$ degrees, where P is the number of phases, and N is the number of rotor teeth. In the case of Fig 1, this gives a step length of 15° indicating that a 24-step sequence is needed to complete one motor revolution.

Hybrid Stepper Motors

In practice the most popular variety of stepper motor is the *hybrid* type, which gives the same type of stepping action as in Fig. 1 but differs in details of construction and operation (for example, its rotor houses a permanent magnet and energising flux flows parallel to the shaft axis). Usually, these motors have four phases or coil windings, which may be available via eight independent terminals as shown in Fig. 2a, or via two sets of triple terminals, as shown in Fig. 2b. The phases are usually designed for unipolar drive and must be connected in the correct polarity.

Figure 3 shows the basic way of transistor driving a normal 4-phase hybrid stepper motor at its designated voltage rating. Table 1 at the end of the

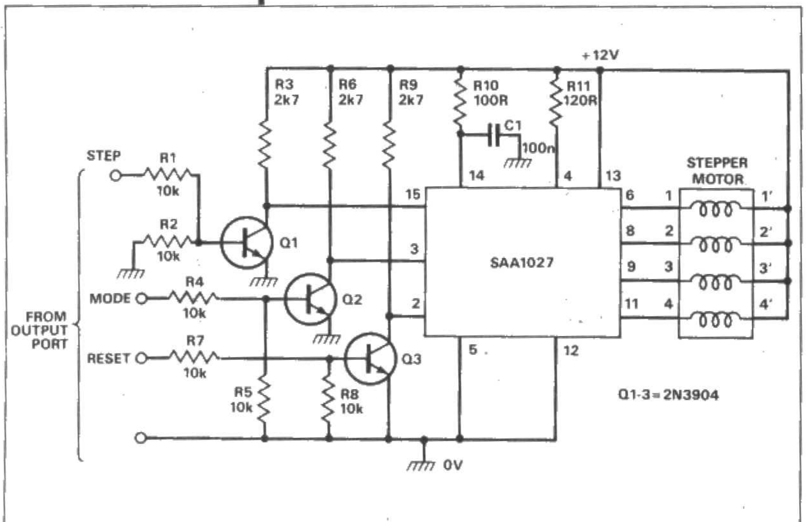

Fig. 7 Stepper motor to microprocessor interface

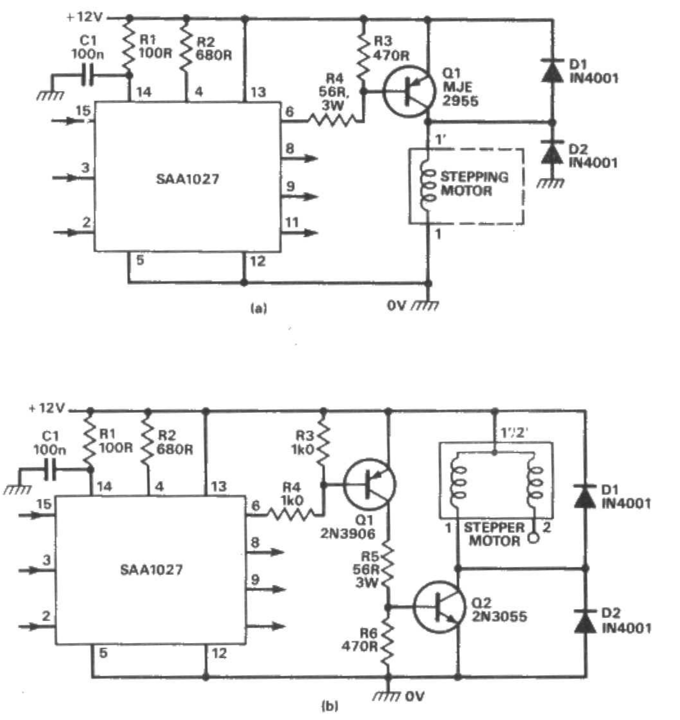

Fig. 8 Two circuits to boost the drive current (a) using independent phase windings and (b) coupled phase windings

article shows the usual full-step switching sequence. Note that the motor can be repeatedly stepped or rotated clockwise by repeating the 1-2-3-4 sequence or anticlockwise by repeating the 4-3-2-1 sequence, and that in each step two phases are energised at the same time but phases 1 and 2 or 3 and 4 are never both on at the same time.

A useful feature of the 4-phase hybrid motor is that it can also be driven in the 'half step' mode, the rotor advancing only a half step angle at a time, using a mixture of single and dual phase switching, as shown in Table 2.

A 4-phase hybrid motor can be operated from a DC supply greater than its designated voltage rating by wiring suitable dropper resistors in series with its phases. Since phases 1 and 2 or 3 and 4 are never both on at the same time, each of these pairs of phases can share a single dropper resistor, R, at each of the points X and Y in Fig. 3. Thus, a 6V 6R motor (1A per phase) can be operated via a 12V supply by giving each resistor a 6R 6W rating.

The SAA1027 Driver IC

Several dedicated 4-phase stepper motor driver ICs are available and the best known of these is the SAA1027, designed to operate from supplies in the 9.5V to 18V range and to give full-stepping 4-phase motor operation at total output drive currents up to about 500mA.

Figure 4 shows the outline and pin notations of the SAA1027, Fig. 5 shows its internal block diagram and basic application circuit. Internally the IC has three buffered inputs which are used to control a synchronous 2-bit (4-state) up/down counter, its output fed to a code converter which then uses its four outputs to control (via suitable driver circuitry) four transistor output stages. Each of these operates in the open-collector mode with motor back-emf damage protection provided via an internal collector-to-pin-13 diode.

Note that the IC has two sets of supply rail pins, one set (pins 13 and 12) feeding the high-current circuitry and the other (pins 14 and 5) feeding the low current sections. In use, pins 5 and 12 are grounded and the positive (usually 12V) rail is fed directly to pin 13 and via decoupling components R1-C1 to pin 14. The positive rail must also be fed to pin 4 via R_x, which determines the maximum drive current capacity of the four output transistors. The appropriate value of R_x is given by

$$R_x = (4E/I) - 60$$

where E is the supply voltage and I is the desired maximum motor phase current. Thus when using a 12V supply, R_x needs values of 420R, 180R, or 78R for maximum output currents of 100mA, 200mA or 350mA respectively.

The SAA1027 IC has three input control terminals, noted COUNT, MODE and RESET. RESET is normally biased high, letting the IC's outputs change state each time the COUNT terminal transitions from the low to the high stage, as shown in Table 3.

The sequence repeats at 4-step intervals but can be reset to zero at any time by pulling RESET low. The sequence repeats in one direction (normally clockwise motor rotation) when the MODE input is low and in the other (normally anticlockwise) when high.

Figure 6 shows a practical drive/test circuit that can be used to drive hybrid 4-phase stepper motors with current rating up to about 300mA. The motor can be manually sequenced one step at a time via SW3 (which is effectively 'debounced' via R4-C5), or automatically via the 555/7555 astable oscillator by moving SW2 to either the STEP or FREE-RUN position. The motor direction is controlled via SW4 and the stepping sequence reset via SW5.

The operating speed of the free-running astable circuit is widely variable via RV1 and SW1. In the SLOW (1) range, the astable frequency is variable from below 5Hz to about 68Hz. On a 48-step (7.5° step angle) motor this corresponds to a speed range of 6-85rpm. Switch SW1 ranges 2 and 3 give frequency ranges that are 10 and 100 times greater than this respectively and the circuit thus gives a total speed control range of 6rpm to 8500rpm on a 48-step motor.

Circuit Variations

The basic Fig. 6 circuit can be varied in several ways. Figure 7 shows how it can be driven via a computer or microprocessor output port with terminal voltages that are below 1V in the logic 0 state and above 3.5V in the logic 1 state. Note that this circuitry reverses the normal polarity of the input control signals. Thus the

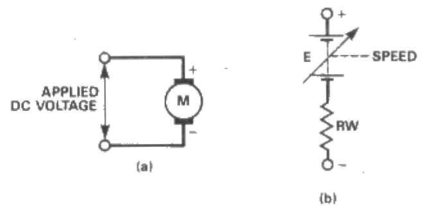

Fig. 9 Symbol and equivalent circuit of a DC permanent magnet motor

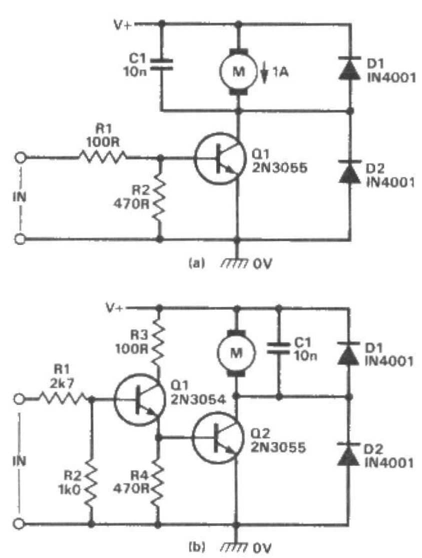

Fig. 10 On/off motor control using transistor switching and (b) increased sensitivity

STEP input is pulsed by a high-to-low transition, the stepping sequence is RESET by a high input, and a low MODE input gives forward motor rotation and a high input gives reverse rotation.

Figures 6 and 7 are designed to give maximum output drive currents up to about 300mA. If desired these outputs can be boosted up to about 5A by using the two circuits of Fig. 8, which each show the additional circuitry needed to drive one of the four output phases of the stepper motor. Four such driver stages are needed per motor. The Fig. 8a circuit can be used to drive motors with fully independent phase windings and the Fig. 8b design can be used in cases where two windings share a common supply terminal. In both cases D1 and D2 are used to damp the motor back-emfs.

Magnet Commutator Motor Basics

Leaving the stepper motor, let's move on to the most widely used DC motor — the permanent magnet type with a commutator that rotates when the motor is powered from an appropriate DC voltage. Figure 9 shows the symbol and simplified equivalent circuit of this type of motor.

The basic action of this motor is such that an applied DC voltage forces current through its sets of armature windings (via commutator segments and pick-up brushes) and generates electromagnetic fields that react with those of fixed stator magnets in such a way that the armature is forced to rotate. As it rotates, its interacting fields make it generate a back-emf that is directly proportional to the armature speed and opposes the applied DC voltage, thus giving the equivalent circuit of Fig 1b, in which R_w represents the total resistance of the armature windings, and E represents the speed-dependent back-emf.

The major points to note about this kind of motor are as follows:

- When the motor is loaded by a fixed amount, its speed is directly proportional to supply voltage.
- When the motor is powered from a fixed DC supply, its running current is directly proportional to the amount of armature loading.
- The motor's *effective* applied voltage equals the applied DC voltage minus the speed-dependent back-emf. Consequently when it is powered from a fixed voltage, motor speed tends to self-regulate, since any increase in loading tends to slow the armature, thus reducing the back-emf and increasing the *effective* applied voltage and so on.
- The motor current is greatest when the armature is stalled and the back-emf is zero. It then equals V/R_w (where V is the supply voltage). This state naturally occurs under 'start' conditions.
- The direction of armature rotation can be reversed by reversing the motor's supply connections.

The main applications of electronic power control to DC motors of this type are in on/off switching and direction control and in variable speed control and improved speed self-regulation (subjects that are dealt with next month).

On/Off Switching

A DC motor can be turned on and off by wiring a control switch between motor and power supply. This switch can be an ordinary electromechanical type (or a pair of relay contacts), or a switching transistor as in Fig. 10a. Here the motor is off when the input is low and is on when the input is high. Note here that diodes D1 and D2 are used to damp the motor's back-emf, that C1 limits unwanted RFI, and R1 limits Q1's base current to about 52mA with 6V input. Under this condition Q1 provides a maximum motor current of about one amp.

In the above circuit, Q1's 52mA base current is provided via the external drive circuitry. If desired, the drive current can be reduced to a mere 2mA or so by adding a buffer transistor as shown in Fig. 10b, where R3 limits Q1's base current to a safe value.

Direction Control

The rotational direction of a permanent magnet DC motor can be reversed by simply reversing the polarity of its supply connections. If the motor is powered via dual (split) supplies, this can be achieved via a single-pole switch connected as in Fig. 11a or via transistor-aided switching by using the circuit of Fig. 11b.

Transistors Q1 and Q3 are biased on and Q2 and Q4 are cut off (with Q2's base-emitter junction reverse biased) when SW1 is set to the *forward* position, and Q2 and Q4 are biased on and Q1 and Q3 are cut off (with Q1's base-emitter junction reverse biased) when SW1 is set to the *reverse* position. Note that if this circuit is used with supply values greater than 12V, diodes must be wired in series with the Q1 and Q2 base-emitter junctions, to protect them against breakdown when reverse biased.

The circuit of Fig. 11b uses double-ended input switching and this makes it difficult to replace SW1 with electronic control circuitry in 'interfacing' applications. Fig. 11c shows how the design can be modified to give single-ended input switching control, making it easy to replace SW1 with electronic switching. In this circuit, Q1 and Q3 are biased on and Q2 and Q4 are cut off when SW1 is set to the *forward* position. Q2 and Q4 are on and Q1 and Q3 are off when SW1 is set to *reverse*.

Direction Control, Using Single-ended Supplies

If a DC motor is powered from a single-ended supply, its direction can be controlled via a double-pole switch

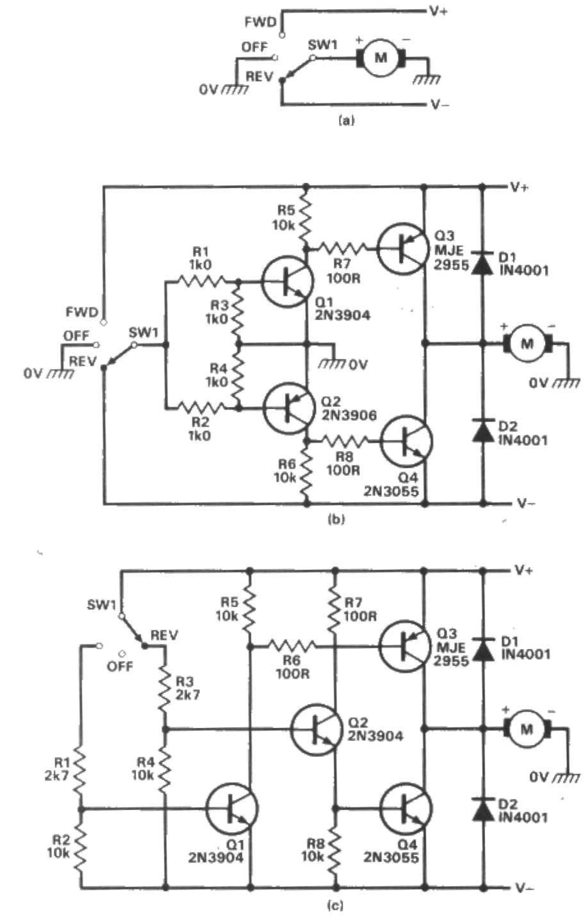

Fig. 11 Motor direction control using (a) simple switching (b) transistor switching and (c) transistor switching with single-ended input

wired as in Fig. 12a, or via a bridge-wired set of transistors connected in the basic form shown in Fig. 12b. Here Q1 and Q4 are turned on and Q2 and Q3 are off when SW1 is set *forward*, and Q2 and Q3 are on and Q1 and Q4 are off when SW1 is set to *reverse*. Diodes D1-4 are used to protect the circuit against possible damage from motor back-emfs, and so on.

Figure 12c shows how the circuit can be modified to give alternative switching control via independent forward/reverse (SW1) and on/off (SW2) switches. A very important point to note about this configuration is that it causes Q1 or Q2 to be turned on at all times, with the on/off action being applied via Q3 or Q4. The motor currents can collapse very rapidly (via the Q1-D2 or Q2-D1 loop) when the circuit is switched off. This so-called 'flywheel' action is vital if SW2 is replaced by a pulse-width modulated (PWM) electronic switch, enabling the motor speed to be electronically controlled (this technique will be described next month).

A weakness of the simple Fig. 12b circuit is that it uses fairly high base drive currents, which must be supplied via the switching circuitry. Fig. 13a shows a more sensitive version of the circuit, which needs input control currents (to the A, B, C, D terminals) of only a few milliamps.

This can be controlled manually via a pair of switches by using the connections of Fig. 13b. It can also be controlled electronically as in Fig. 13c, in which a 4052B CMOS IC is used as a ganged 2-pole 4-way bilateral switch that can be controlled via logic-0 or logic-1 signals applied to its 'A' or 'B' input pins, to give independent forward/reverse and on/off (or PWM speed control) actions. Note that both of these circuits are configured to give the 'flywheel' type of switching action already described.

To conclude this month's article, Table 4 illustrates the above point by showing the truth table that occurs when the Fig. 13a and 13c circuits are interconnected.

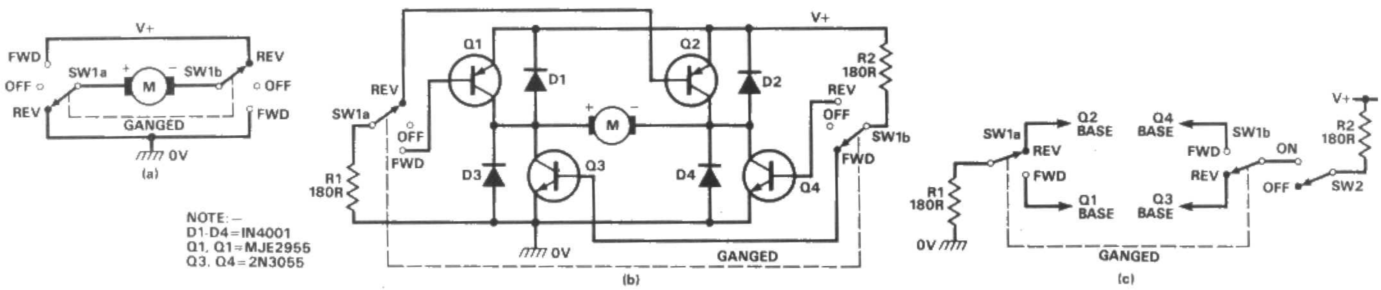

Fig. 12 Motor direction control with single-ended supply (a) simple switching (b) transistor switching (c) separate on/off switching

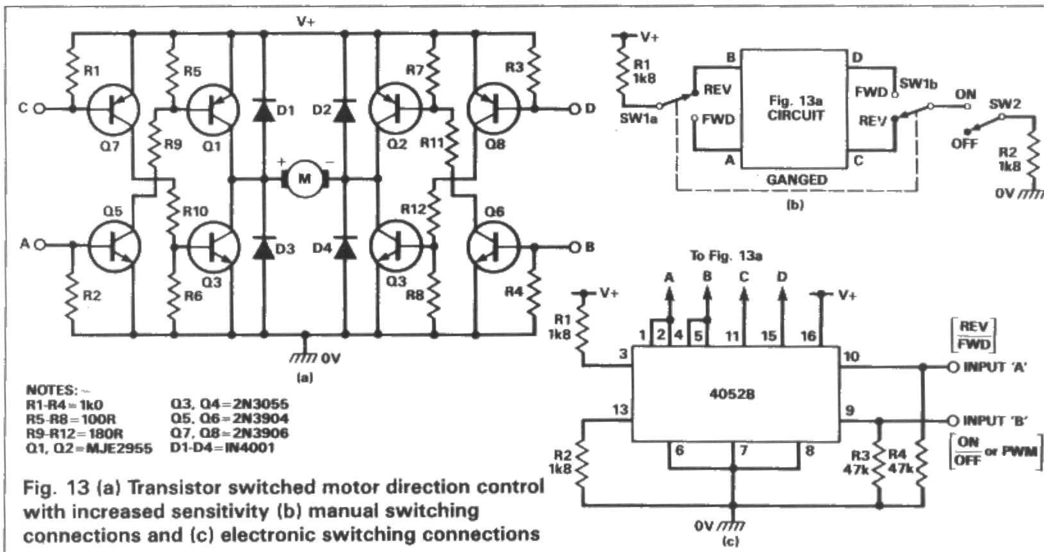

Fig. 13 (a) Transistor switched motor direction control with increased sensitivity (b) manual switching connections and (c) electronic switching connections

DC MOTOR TABLES

STEP No.	Q1	Q2	Q3	Q4
0	ON	OFF	ON	OFF
1	OFF	ON	ON	OFF
2	OFF	ON	OFF	ON
3	ON	OFF	OFF	ON
4	ON	OFF	ON	OFF
5	OFF	ON	ON	OFF

Table 1 Full step sequencing of the circuit in Fig. 3

COUNTING SEQUENCE	MODE-LOW				MODE-HIGH			
	Q1	Q2	Q3	Q4	Q1	Q2	Q3	Q4
0	ON	OFF	ON	OFF	ON	OFF	ON	OFF
1	OFF	ON	ON	OFF	ON	OFF	OFF	ON
2	OFF	ON	OFF	ON	OFF	ON	OFF	ON
3	ON	OFF	OFF	ON	OFF	ON	ON	OFF
0	ON	OFF	ON	OFF	ON	OFF	ON	OFF
RESET LOW	ON	OFF	ON	OFF	ON	OFF	ON	OFF

Table 3 SAA1027 output sequencing table

STEP No.	Q1	Q2	Q3	Q4
0	ON	OFF	ON	OFF
1	ON	OFF	OFF	OFF
2	ON	OFF	OFF	ON
3	OFF	OFF	OFF	ON
4	OFF	ON	OFF	ON
5	OFF	ON	OFF	OFF
6	OFF	ON	ON	OFF
7	OFF	OFF	ON	OFF
8	ON	OFF	ON	OFF
9	ON	OFF	OFF	OFF

←Clockwise← →Anticlockwise→

Table 2 Half step sequencing

FIGURE 13c CIRCUIT STATES						FIGURE 13a TRANSISTOR STATES			
INPUTS		OUTPUTS							
'A'(R/F)	'B'(ON/OFF)	A	B	C	D	Q1	Q2	Q3	Q4
0	0	1	X	X	X	ON	X	X	X
0	1	1	X	X	0	ON	X	X	ON
1	0	X	1	X	X	X	ON	X	X
1	1	X	1	0	X	X	ON	ON	X

Table 4 Truth tables of Figs 13a and 13c when interconnected

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NEON CIRCUITS

Paul Chappell's neon lights alive with a selection of summer circuits to get your workbench glowing

Let's face it, nobody want to sit slaving over a hot soldering iron for hours on end when the sun is shining. So here are some projects to amuse you for a few minutes during the odd rain shower. None should take more than twenty minutes to build, so by the time the sun comes out again — you've finished. All are based on that very underrated component, the neon bulb.

Here's a very easy one to start with: a neon flasher. It has lots of uses like . . . um . . . a beacon to stop low flying owls crashing into your roof at night, for instance. My house hasn't suffered any owl damage at all since I installed mine.

Otherwise, it can be a warning lamp to attract your attention to a fault condition (Fig. 2 shows a blown fuse indicator), or use it just as an alternative to a mains-on lamp: if you get fed up with it staring at you, make it wink instead.

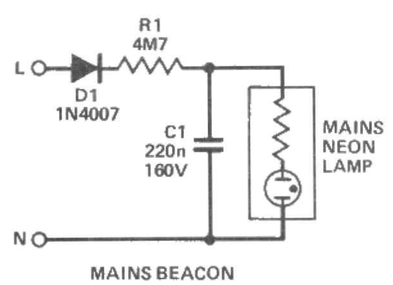

Fig. 1 Circuit diagram for a mains beacon

The circuit is shown in Fig. 1. On positive half-cycles of the mains, C1 charges up via D1 and R1. Eventually the voltage across C1 will be high enough for the neon to strike and glow. The current taken by the neon discharges C1 until the voltage is too low for the neon to continue glowing. It turns off, C1 begins to charge and the cycle repeats for ever more.

The important characteristic of the neon bulb in this circuit, and in all the others for that matter, is that it takes a higher voltage to make the neon start to conduct than it does to keep it going once it's started. It may strike at 80V but won't turn off again until the voltage sinks below, say, 50V. That's what makes it possible to use it as an oscillator.

Fig. 2 shows the circuit connected as a blown fuse indicator. As long as the fuse is intact it shorts out the flasher circuit. Once the fuse blows, the flasher takes current via the transformer — but such a small current that the protected circuit won't be powered up at all. Construction details are below (if you really need them!). For the beacon or mains indicator, just connect the circuit directly between L and N of the mains.

Fault finding couldn't be simpler. If anything is wrong, the lamp will either stay on all the time which means that R1 is too small, or will not come on at all which means that C1 is too leaky. With the DC supply to the neon, the glow will only be seen around one of the internal electrodes — this is an unavoidable characteristic of the circuit and doesn't indicate a fault. To increase flash rate, reduce the value of C1. To make it slower, increase the value of C1.

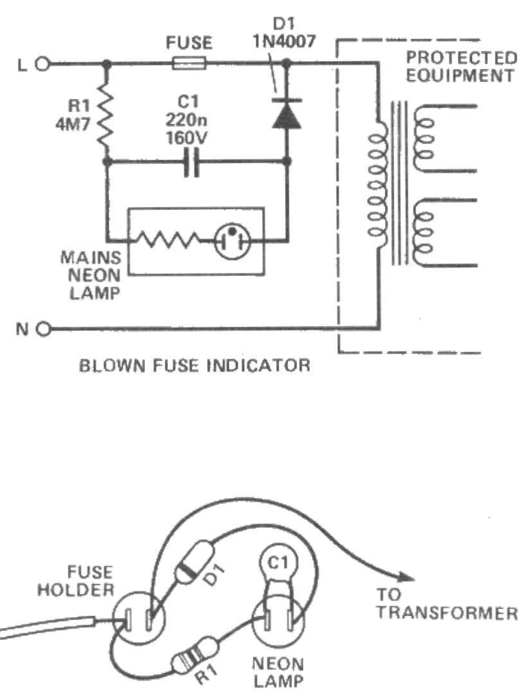

Fig. 2 Circuit diagram and construction for a blown fuse indicator

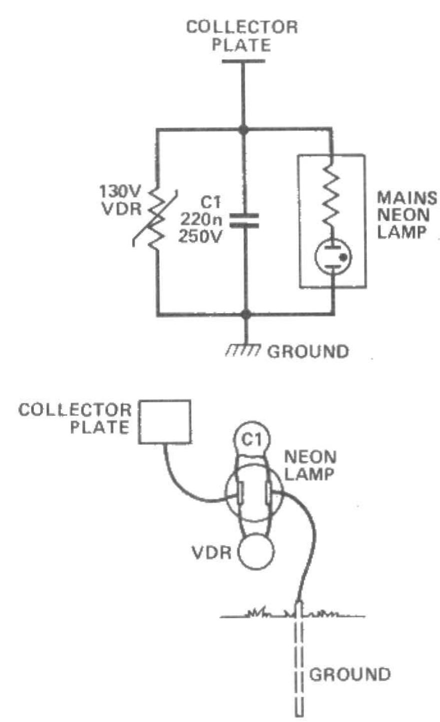

Fig. 3 Circuit and construction for an electrical activity indicator

Anyone who is enthusiastic about Amateur Radio or home weather forecasting will find the next little circuit useful. It gives an indication of electrical activity in the air, which has a great effect on the quality or even the possibility of picking up distant radio signals. During quiet periods the neon will hardly flash at all. During a thunderstorm it will be on almost continuously. Degrees of electrical activity between these two extremes will be shown by variations in the flash rate.

The circuit and construction details for this project are shown in Fig. 3. The VDR is included to sink surplus current which might otherwise damage the neon during times of intense activity. The collecting plate should be insulated — a piece of PCB material will be fine — and fixed high on the outside wall of the house. Not on the roof, though. It's not supposed to be a lightning conductor!

The wire connecting the plate to the circuit should be well insulated but thin enough to fuse immediately if a lightning strike should ever occur close by. The ground should be exactly that — a metal rod driven two feet into the ground. An offcut of copper pipe will do nicely.

In Fig. 4 there's a coin toss circuit, or a make-up-your-mind machine for the indecisive. The circuit is similar in principle to the two transistor multivibrator that will be familiar to anyone who was around in those far off days when ICs were a rarity and transistors

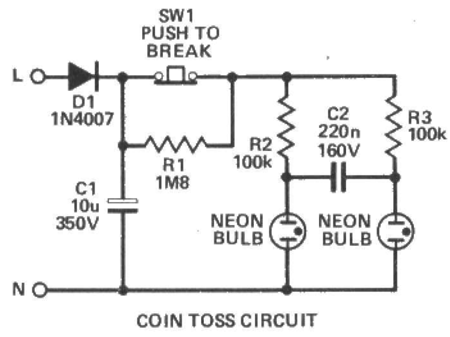

Fig. 4 Circuit for a coin toss decision maker

the main active components (Fig. 5). Suppose the left-hand neon of Fig. 4 is on and the right one isn't. C2 will charge via R3 until it eventually reaches the point where the right hand neon will strike. If we stick to the 80V and 50V I mentioned earlier as fairly typical for striking and extinguishing a neon, then just at the

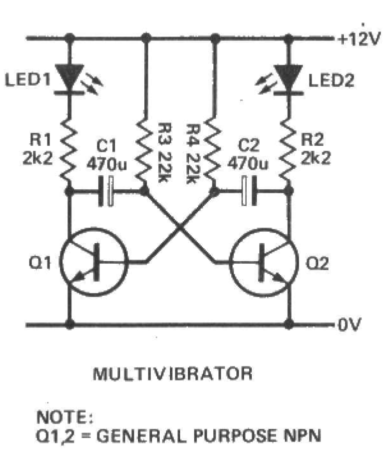

Fig. 5 Transistor multivibrator circuit

moment the right hand neon is on the point of striking, the right hand side of C2 will be at 80V and the left hand side at 50V. As soon as the right hand neon strikes, the voltage at the right hand plate of C2 will drop to 50V and the left hand plate to 20V, so now the right hand neon will be on and the left hand one off. Now the right hand plate is held at 50V. The left hand plate of C2 will charge via R2, the left hand neon will eventually strike, turning off the right hand one at the same time.

And so it continues, with first one neon and then the other lighting. With the component values shown, the flashing will be almost fast enough to be taken for continuous illumination of both bulbs, although they will flicker.

Push SW1 and an extra resistor is introduced into the circuit. The value of R1 is chosen so that the conducting neon drawing current through it will drop enough voltage that the top of R2 and R3 will be below 80V, so the neon that isn't lit when the switch is pressed can never light at all. In other words, the flashing is frozen, with whichever neon was on when the switch was pressed staying on. Label the neons 'yes' and 'no' and press for an instant decision!

The value of R1 may need a little adjustment by trial and error to suit the particular neon lamps you use. If too small, the neons won't stop flashing. If too large, the lit neon will be very dim when the switch is pressed.

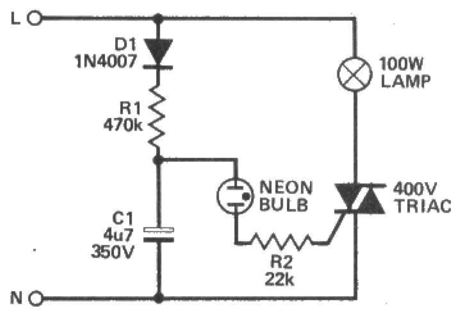

Fig. 6 Circuit with neon as trigger

Yet another use for a neon bulb is as a trigger for triacs and SCRs. Figure 6 shows the general idea. This circuit will flash a mains lamp if you have an urge to do such a thing. Figure 7 shows a circuit for an electric fence energiser. These are more by way of tech tips than projects, although Fig. 6 will work perfectly well as it stands.

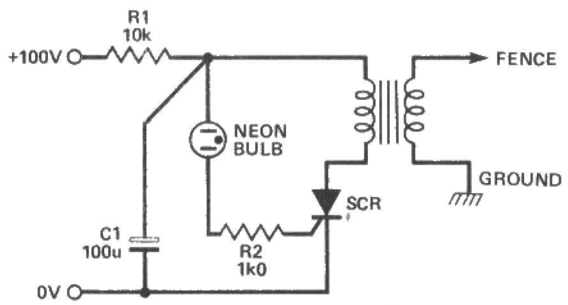

Fig. 7 Circuit for an electric fence energiser

When you're experimenting with any of these circuits, do be careful — none are isolated from the mains. Also bear in mind that the caps can retain enough charge to give a shock even when the mains is disconnected, so discharge each one through a 100R resistor before making any mods to the circuit.

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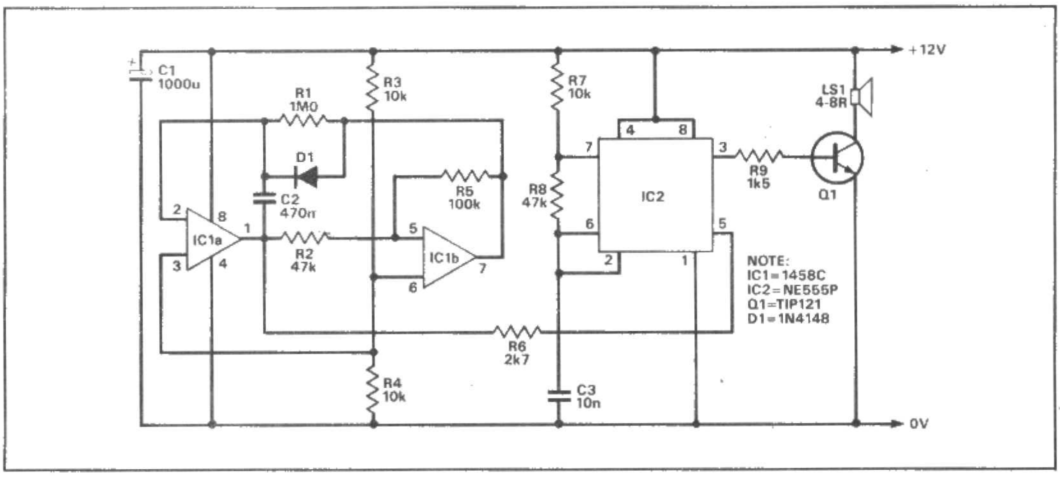


Robert Penfold presents a package of circuits to infuriate intruders

Alarm Generator

Many alarm systems still rely on the standard bell alarm generators, remaining as effective as they ever were. Electronic alarm generators are increasingly popular though, and represent an easy project for the home constructor. A simple signal having a single unmodulated tone is not very effective as it is easily masked by other sounds — not sounding much like an alarm signal it runs the risk of being ignored even if someone should hear it! This alarm generator relies on frequency modulation to give a very effective sound and this is probably the best type of modulation for alarm applications.

IC2 generates the basic tone signal as a 555 timer in the standard astable configuration. The operating frequency is in the region of 1kHz and the often unused pin 5 of the 555 is brought into play in this circuit. Normally timing capacitor C3 repeatedly charges to two thirds of the supply voltage and discharges to one third of the supply potential. Pin 5 provides access to the potential divider that sets the two thirds of V+ threshold level and permits this voltage to be modulated. By raising this voltage, it takes longer for C3 to charge and discharge from the threshold level, the operating frequency being reduced. A lower threshold voltage gives reduced charge and discharge times and a higher output frequency. IC2 can therefore be frequency modulated via a modulation voltage applied to pin 5.

The modulation signal is generated by IC1 connected in what is almost the standard square/triangle oscillator configuration operating at about 5Hz. IC1 functions as the Miller integrator while IC1b provides the Schmitt trigger action. The circuit differs from the standard configuration in that D1 is included across timing resistor R1 so that the charge time of C2 is greatly shortened — a sawtooth waveform is obtained instead of a triangular type.

In terms of the audio output signal, this gives a tone which is swept downwards in frequency, switched up to a higher frequency, swept downwards again and so on. The effect obtained is easily modified. To obtain an upwards sweep it is merely necessary to reverse the polarity of D1. For a smooth upwards and downwards sweep D1 should be removed altogether (this also halves the modulation frequency). The modulation frequency is inversely proportional to the value of C2. The output frequency range is controlled in the same way by C3.

The output power available from IC2 is only quite low and is totally inadequate for most purposes. The loudspeaker is therefore driven via an emitter following Darlington power device Q1 — readily able to provide output currents of a few amps. The output power is a few watts into a 4-8R impedance loudspeaker using a 12V supply. If high output power is important, use a 4R speaker and raise the supply voltage to 15V. Even using a 12V supply and 8R speaker the sound is pretty loud unless an inefficient speaker is used. In this application efficiency is of greater importance than audio quality and it is also important that the loudspeaker is rated to handle about 8W rms or more. Miniature types are likely to be destroyed after a few seconds of operation! As Q1 is used in a switching mode it does not dissipate much power but it is advisable to fit it with a small finned heatsink. The current consumption of the circuit is about 800µA using an 8R loudspeaker, or around double this figure for a 4R type.

Door Alarm

While there is no real substitute for a comprehensive burglar alarm system, many people are simply not prepared to go to the trouble or expense of having such a system installed on their premises.

A simple form of alarm is considerably better than nothing — the most popular being the various types of window and door alarms. These vary in sophistication but most rely on some form of mechanical or magnetic switch to trigger a simple electronic alarm generator circuit. One way of handling the sensor part of such an alarm is to have a reed switch fitted in the main unit, mounted on the door or window. The actuating magnet is fitted on the door or window frame, with everything arranged so that the magnet activates the switch when the door or window is closed. In some cases it might be more convenient to have the main unit mounted on the frame and the magnet on the door or window.

There are alternative types of switch — one possibility is to use a micro-switch. One of these could be rather awkward to use in a simple stand-alone alarm though and my preferred method is to use some form of vibration switch. This could be a mercury switch mounted very close to the angle at which it is activated, or a pendulum type mechanism — when

the unit is moved the pendulum swings and comes into contact with a circular electrode, completing a circuit in the process. Ready-made switches of this type can be obtained, otherwise it is not too difficult to improvise a DIY version.

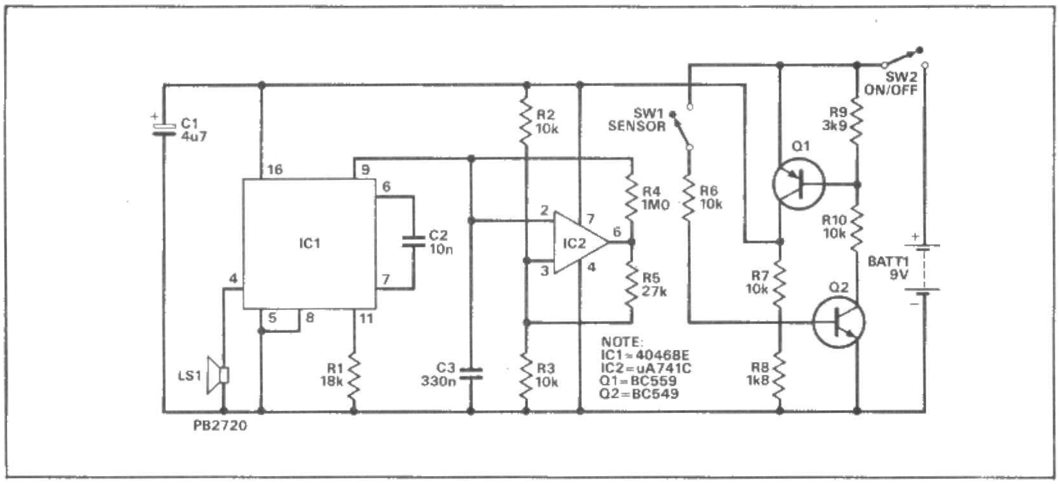
The electronics of such a simple alarm can be nothing more than a low power alarm generator circuit. However, it is advisable to include a latch circuit. Remember that the sensor switch may be activated only momentarily and that without the aid of a latch the alarm generator may produce nothing more than one or two very brief 'squeaks'.

In this alarm unit the latching action is provided by Q1 and Q2 connected in a sort of pseudo thyristor arrangement. When sensor switch SW1 is closed, Q2 is biased into conduction and its collector current biases Q1 into conduction. Feedback through R7 results in both transistors being held in the *on* state even if SW1 should open again. Note that SW1 must provide a normally closed action.

The alarm generator circuit is based on a 4046BE CMOS low power phase locked loop (IC1). In fact it is only the VCO section of the device that is utilised in this circuit — the phase detectors and the other stages are left unused. The VCO generates the basic alarm signal and drives a ceramic resonator (LS1). The latter provides a very high level of efficiency and can provide a very loud alarm signal from the very limited output current of IC1.

In order to obtain this efficiency the output frequency of IC1 must be in the region of 1-5kHz and it is swept up and down over this range by IC2 which operates as a standard operational amplifier astable circuit. There is a squarewave output signal from IC1 but in this case it is the non-linear triangular waveform across timing capacitor C3 that is used as the modulation signal. The frequency is about 5Hz.

The unit should provide many months of operation in the standby mode as the current consumption in this state is only the leakage current that flows through Q1 and Q2 (typically under 1μA). The current consumption when the unit is activated is only a few milliamps and a small (PP3 size) battery should suffice as the power source. Ideally SW2 should be a key-switch so that there is no easy way for an intruder to switch the unit off once it has been triggered. The unit is reset by switching it off for a second or two and then switching it back on again.



Dopper Shift Alarm

No doubt ETI readers are all familiar with the doppler shift effect which, for example, gives the apparent change in the engine pitch of a passing car. Doppler shifts are now used to good effect in a number of practical applications including such diverse tasks as navigation systems and burglar alarms. In this second category there are radar systems (using microwave techniques) and ultrasonic systems. Radar systems have their advantages but for the home constructor there can be licensing problems plus they are relatively expensive. The system described here is a low cost but very sensitive ultrasonic system.

A unit of this type transmits an ultrasonic signal and picks up any reflected signals. Normally the reflected signals will all be at the same frequency (the transmission frequency). However, if something (or someone) moves around in the area covered by the alarm, the reflected signal is shifted slightly in frequency due to the Doppler effect. Some of the signal picked up by the unit is at the transmission frequency and some is at a slightly different frequency. The two signals interact to give what is effectively an amplitude modulated signal, with the modulation at a frequency equal to the difference between the two. This is the same effect that causes a tone to be heard when AM radio is tuned to two stations on marginally different frequencies. The alarm operates by demodulating the received signal and using any audio output signal to trigger a relay driver circuit. In fact the demodulated signals will often be at sub-audio frequencies and the relevant stages of the unit must be designed to cope with these very low frequencies.

This ultrasonic alarm provides virtually full coverage of anything but the largest of rooms in a house. Although systems of this type generally have a few blind spots, these are generally inadequate to permit someone to enter and leave a room without triggering the alarm. Apart from anything else, someone moving around the room will usually produce a certain amount of turbulence that will spread across the room and trigger the alarm.

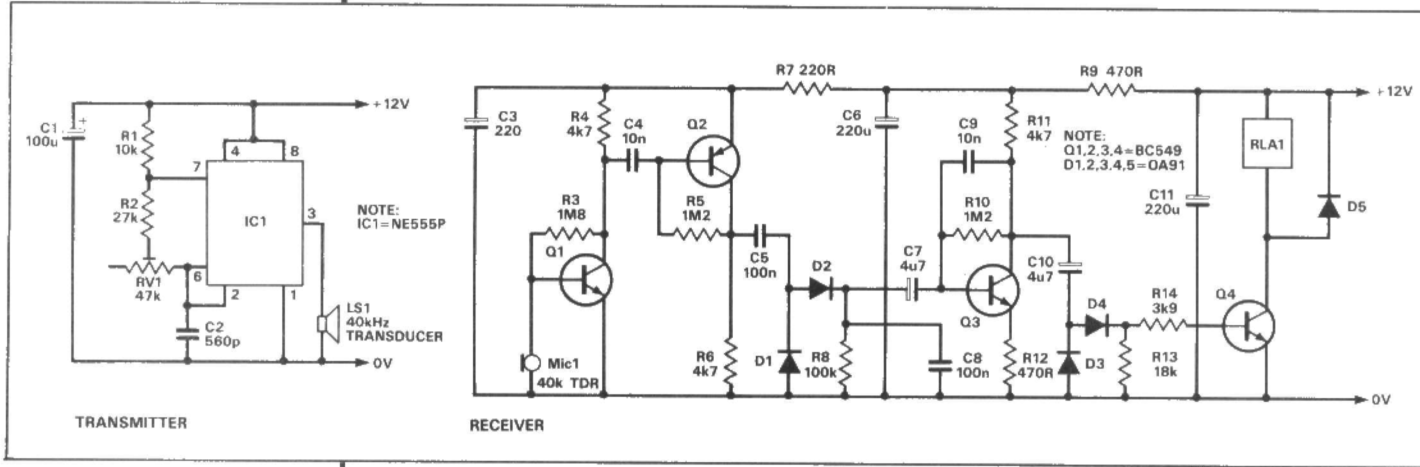
The transmitter circuit is a standard 555 astable driving a 40kHz ultrasonic transducer. RV1 is used to adjust the transmitter for the frequency that gives optimum results.

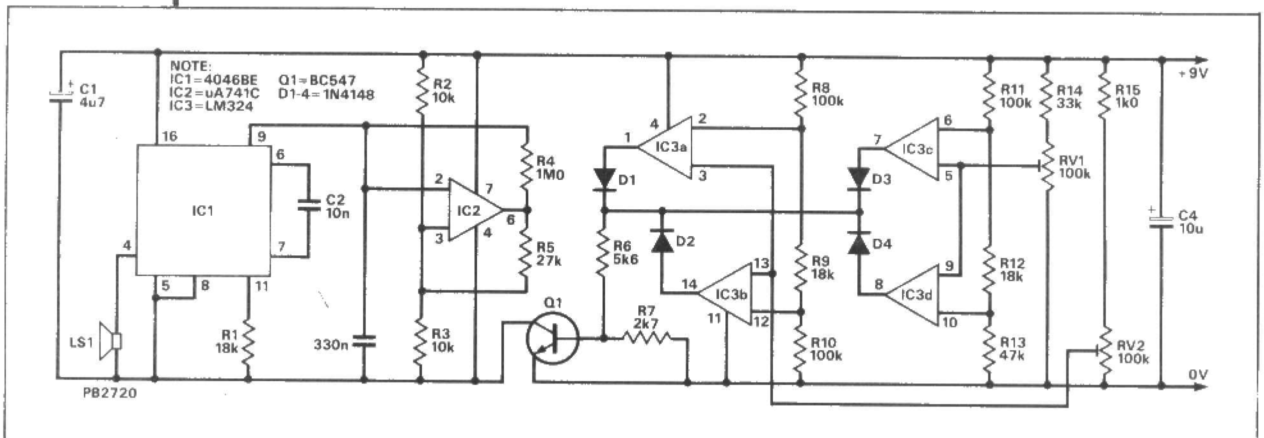
At the receiver another ultrasonic transducer operates as the microphone, its output fed to a two stage high gain amplifier based on Q1 and Q2. The amplified signal is fed to a diode demodulator and then to a common emitter amplifier which provides about 20dB of voltage gain. The output of this stage is fed to a rectifier circuit and then to a relay driver stage. When modulation is present on the received signal, positive going half cycles from D4 will drive Q4 into conduction and activate the relay. Note that the circuit does not provide a latching action and it is assumed that it will be used as part of a standard switch type burglar alarm. In other words, this unit will be

connected in with the window switches, trigger mats and other sensors of the alarm system. The main alarm circuit will then provide latching, exit delays etc.

Any 40kHz ultrasonic transducers should suffice for LS1 and MIC1 but note that with some of these there are specific transmitter and receiver units, while with others the two transducers are identical. Be sure to get them round the right way where necessary. Try to have the transducers mounted at least 150 millimetres apart. RLA1 can be any 12V relay with a coil resistance of about 200R or more and suitable contacts.

Due to the high gain of the receiver circuit some care needs to be taken with the component layout or instability might occur. If suitable test gear is not available, RV1 must be adjusted using trial and error to find the setting that gives optimum pick up and sensitivity. A little experimentation should soon find a good position and aim for the unit. In general, results tend to be best with the unit in a corner or somewhere at the edge of a room, not near the middle. Sensitivity of the circuit can be reduced by increasing R12 and vice versa. However, very high levels of sensitivity can bring problems with false alarms. The slightest draughts or small insects flying around the room can be sufficient to trigger the unit!





Loop Alarm

A loop alarm is a form of alarm that is often used in shops and on market stalls. The basic idea is to have a loop of wire which is threaded through the goods which are to be protected by the alarm-radios and televisions for instance. All the alarm circuit has to do is to sound the alarm if the wire is broken.

In practice this type of alarm is not as good as it could be in that it is not too difficult to bypass the wire so that it can then be cut without activating the alarm. There are ways of making such an alarm more 'crack-proof' though, and the method used in this design is to have a multi-way cable rather than a single wire. Breaking any one of the five leads, or several of them, will activate the alarm. Any cross coupling or other tampering will almost certainly activate the unit.

The loop can be bypassed but all four wires must be individually bypassed (with no accidental short circuits) in order to successfully 'crack' it. In a practical situation there is not likely to be a realistic chance of anyone managing this without being noticed!

The alarm generator section of the unit is exactly the same as the one used in the door alarm circuit described earlier — a high pitched FM tone which gives quite good volume from a high efficiency ceramic resonator.

A pair of window discriminators form the basis of the broken loop sensor part of the circuit. If we consider IC3a and IC3b, these are operational amplifiers both operating here as voltage comparators. They each have one input fed from the potential divider formed by R8, 9 and 10, with IC3a receiving the higher reference voltage and IC3b being supplied with the slightly lower one. The other two units are wired together and fed with a variable voltage from RV2.

If the voltage supplied by RV2 is greater than the higher reference voltage, the output of IC3a goes high and activates the alarm by way of Q1. Similarly, if the voltage from RV2 is below the lower reference voltage, the output of IC3b goes high and activates the alarm. In practice RV2 is adjusted to a voltage between these two levels so that the alarm is not normally activated.

IC3c and IC3d form what is virtually an identical

window discriminator circuit but the 'window' is at a lower range of voltages. RV1 is adjusted for a voltage that is within this range.

RV1 and RV2 are connected to the main circuit via the loop of wire, which should be a quad screened cable having the negative supply rail carried by its outer braiding. The cable does not actually need to be a loop and the two preset potentiometers can be mounted in a separate case some distance away from the main unit if preferred. If any of the connecting leads are cut or short circuited, the input voltages are taken outside the 'windows' and the alarm is switched on. R14 and R15 ensure that the unit can not be defeated by connecting the wires together so that the power supply is short circuited.

The circuit does not incorporate latching but once the cable is cut or otherwise tampered with it is unlikely that there would be any easy way of reversing the process and silencing the alarm. The unit should work properly with a connecting cable of up to at least 10 metres in length.

The standby current consumption of the unit is about 800μA rising to about 5mA when activated. Battery operation is therefore feasible provided a fairly high capacity type is used (such as six HP7 size cells connected in series).

Gas Alarm

There are a surprisingly large number of ways of detecting fire, or a situation where fire or explosion is a real hazard. There are a range of methods which rely on optical sensors to detect smoke, or turbulence in the air caused by the heat of a fire. Some units simply detect an abnormally high temperature, or an unusually rapid increase in temperature.

Another method, and a highly sophisticated one, is to use a form of sensor that detects inflammable gases or vapours in the air. These can usually detect the smoke from a fire without difficulty but they are more than just fire detectors as they can detect inflammable gases and vapours at quite low concentrations. A unit based on one of these sensors can therefore be used as a gas alarm as well. They are particularly valuable for use in caravans and boats, where bottled gas is used. They can detect the risk of explosion due to a gas leak, rather than simply detecting, after the event, that the remains of the boat (or whatever) are on fire.

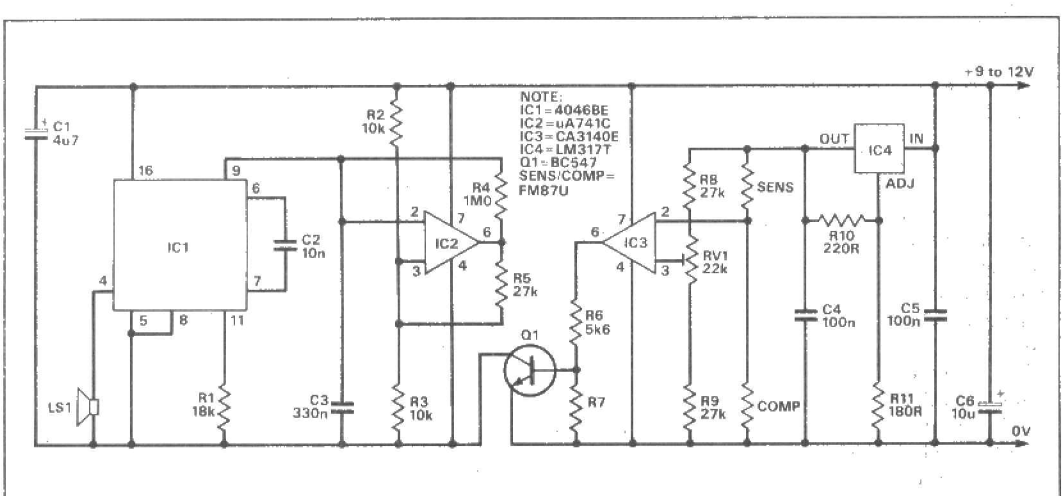
This gas alarm is based on the detector available from Maplin. Actually this is two components — the sensor and a compensating element. The sensor has fine platinum wire coated with oxides and a catalyst. Normally the current passing through the sensor causes it to heat to approximately 350°C but in the presence of suitable gases oxidisation takes place. This produces a rise in temperature which in turn causes the increase in the resistance of the sensor. The compensating element has characteristics which are almost identical to the sensor but in the presence of combustible gases its temperature and resistance are unaffected.

The sensor and compensating element are designed to be used in a bridge circuit. They form one side of the bridge while R8, R9 and RV1 are used in the other section. IC3 operates as a voltage comparator and RV1 is set for a wiper voltage that is marginally lower than the voltage produced by the sensor side of the bridge. The output of IC3 therefore goes low, Q1 is cut off and the alarm circuit is not activated.

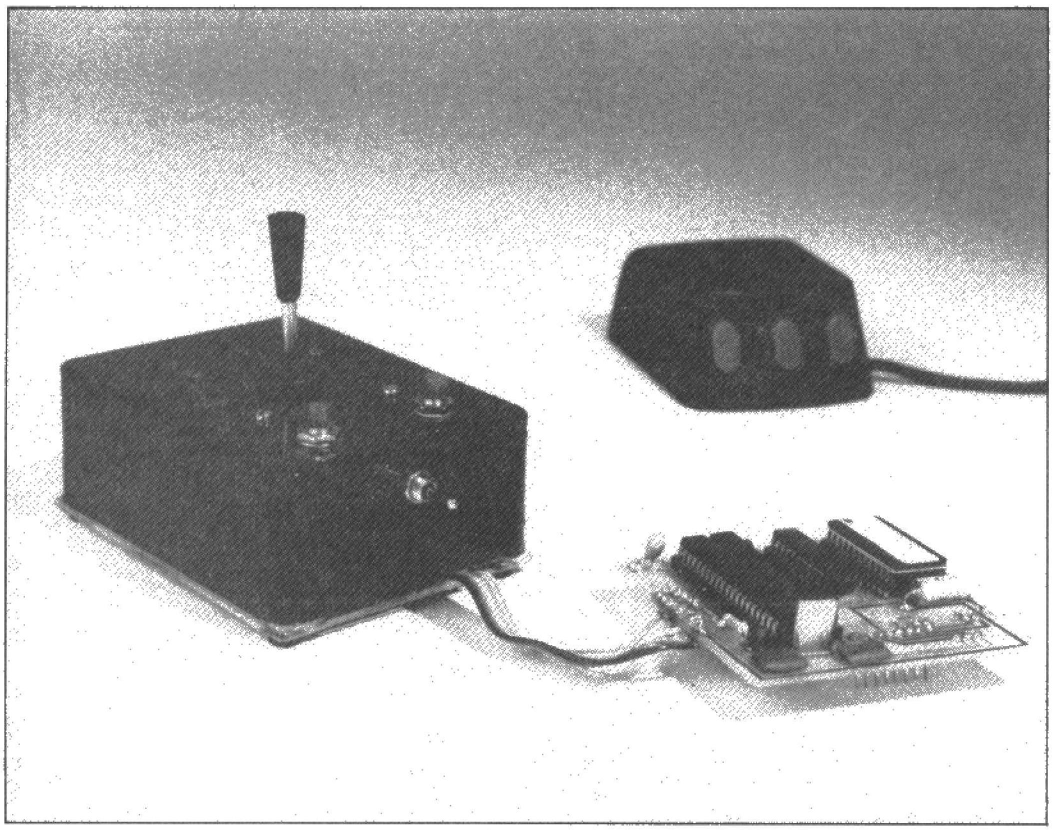
The sensor circuit requires a 2.2V supply and this is provided by a voltage regulator based on IC4. Any drift in the output voltage of IC4 should not adversely affect operation of the circuit as it will affect both sides of the bridge equally. Any environmental changes that affect the sensor will affect the compensating element by an almost identical amount and this will prevent any change in the output voltage from this side of the bridge. Of course, if inflammable gas is detected by the sensor, its resistance rises and the output voltage from this side of the bridge circuit falls. This takes the inverting input of IC3 to a lower voltage than that present at the non-inverting input. The output of IC3 then triggers to the high state, Q1 is switched on and the alarm generator is activated.

It is important to get the sensor and compensator round the right way — the one with a blue spot on the body is the compensating element and the one without the blue spot (it will probably have two or three spots of other colours) is the sensor. There is a change in output voltage from the sensor circuit of only about 20mV when the unit is activated. It is therefore important to carefully adjust RV1 for the lowest possible voltage that does not trigger the alarm.

Note that it takes about 10 seconds for the sensor and compensating element to stabilise after switch-on. The current consumption of the circuit is quite high at about 350 to 400mA. There is no risk of the sensor or compensating element igniting suitable concentrations of inflammable gas as they are both covered by a double mesh of stainless steel. Note that IC4 should be fitted on a medium sized heatsink as it has to dissipate a few watts of power.



PROJECT



DIGITAL JOYSTICK -TO-MOUSE CONVERSION

Richard Grodzik presents a design to control a computer mouse port with a considerably more compact digital joystick

Invariably, all personal computers today are supplied with a mouse and WIMP (window-icon-mouse-pointer) oriented software packages, thus reducing the need for keyboard skills and hopefully producing a friendlier machine.

The mouse of the WIMP acronym is a small opto-electronic device which sits on the desk next to the computer. As the user moves it around the desktop, a pointer on the computer screen mimics its movements.

This gives an easy way to select options from the screen — there is no need to type in strangely-named commands. Many commands are shown by small pictures (icons) on the computer screen with a list of options. The user just points at the required option and pushes a button on the mouse to select.

If we perform an autopsy on a typical mouse (see Fig. 1) we find a ball which sits on two freely rotating steel shafts at right angles to each other. The two shafts drive optical encoders which produce electrical pulses, the number of pulses produced being proportional to the vector quantities derived from the balls' movement. These pulses are shaped and squared by schmitt-triggers and fed to the computer, which translates these signals into pointer movement on the screen. Thus, as the mouse is dragged around the desktop, a pointer on the screen follows its movements.

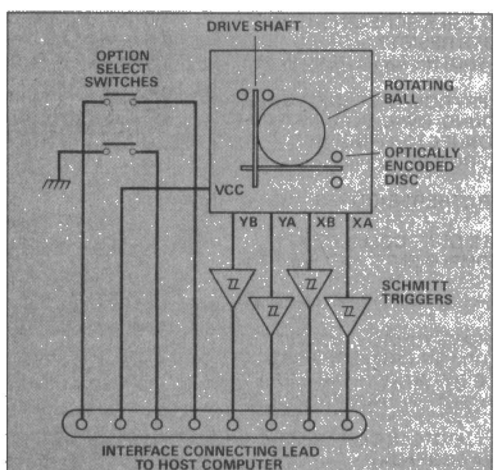


Fig. 1 Block diagram of a standard mouse

The ball mechanism of the mouse is subject to the ingress of dirt as well as wear and tear. Indeed, it has been known for the ball to become mysteriously detached and lost! In addition the proliferation of paperwork associated with computer use often doesn't leave the mouse enough room to move about.

A further consideration to be taken into account is the actual speed of the pointer across the screen. Mouse movement scaling (the relationship between

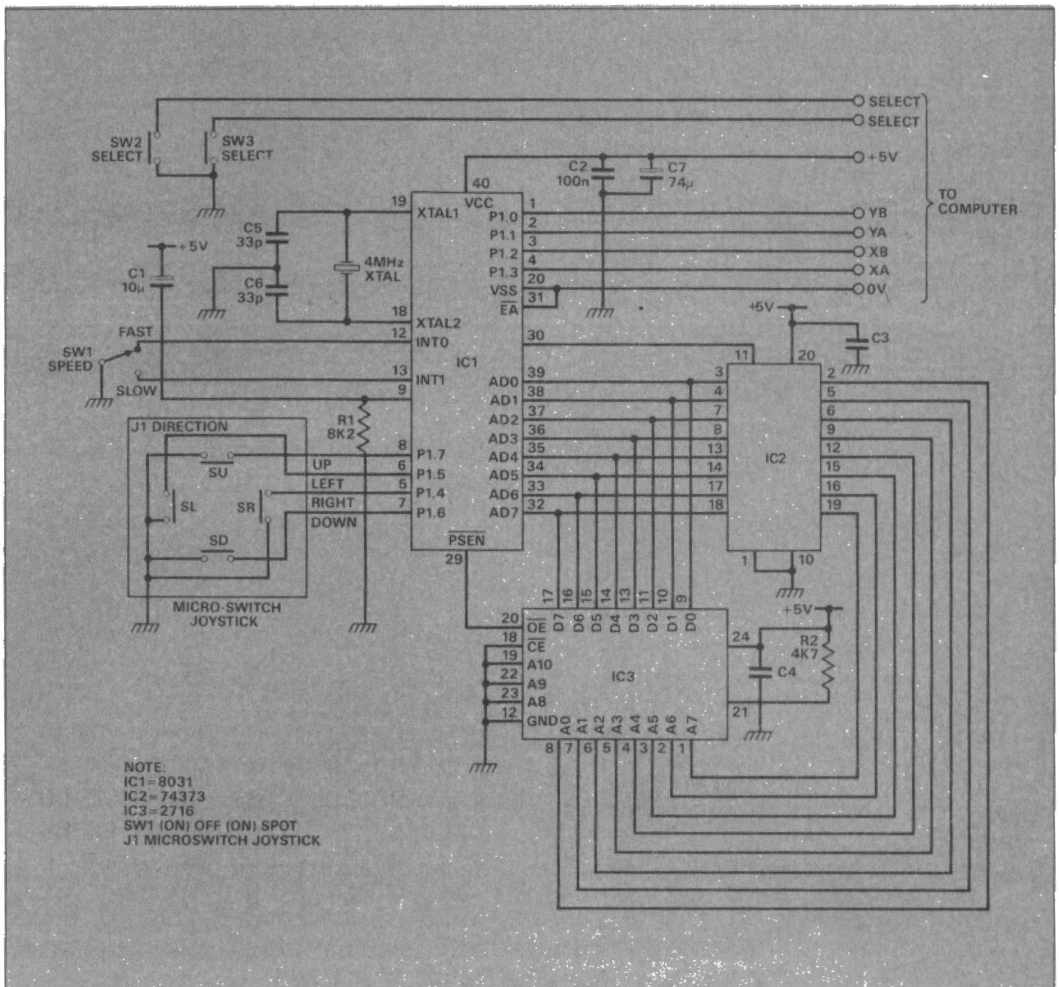


Fig. 2 The circuit diagram of the electronic mouse-on-a-stick

how far you move the mouse and how the pointer moves on the screen) can be re-configured by accessing the operating system and consulting the not-so-user-friendly computer user's manual.

This digital mouse design does away with the opto-electronic mouse philosophy adopted by most manufacturers. The rotating ball mechanism has been replaced by a joystick which has the major advantage that it requires very little desk space. Bin that mouse mat and you've got room for an extra folder or a plate of microwaved chicken orientale and half a lager. A work environment transformed!

And don't baulk at the word joystick. Those cheap plastic nasties supplied for computer games are mostly analogue devices utilising potentiometers and/or switches which connect to the computer's analogue port. The digital joystick recommended here connects into the same digital port as the original mouse and is a direct replacement, so that an old plug and connecting lead from the mouse may be utilised.

The joystick has eight axes of movement, enabling the screen pointer to move in a vertical, horizontal or diagonal direction. Pointer speed or movement scaling is adjustable without recourse to the computer's operating system by means of an spdt switch. Pressing the switch in one direction increases the pointer speed, in the other it decreases the speed — those intergalaxian mind spleens with their turbo spaceprops had better watch out!

The ability to reduce the pointer speed to a slow crawl has proved to be invaluable in design systems where precise pointer positioning is required — drawing a straight line with a conventional mouse can be a tricky thing sometimes.

Construction

Building the unit on the ETI PCB (Fig. 4) is reasonably simple. There are few discrete components. Capacitors C5 and C6 can be laid flat against the topside PCB tracks — this also reduces the risk of a short with the case of the crystal. Sockets are recommended for all the ICs and obviously take precautions with static sensitive devices.

BUYLINES

Electromail (telephone (0536) 204555) can supply suitable joysticks: the small version has order code 332-105, a larger model is 332-111. The PCB is available from the ETI PCB Service (see centre pages).

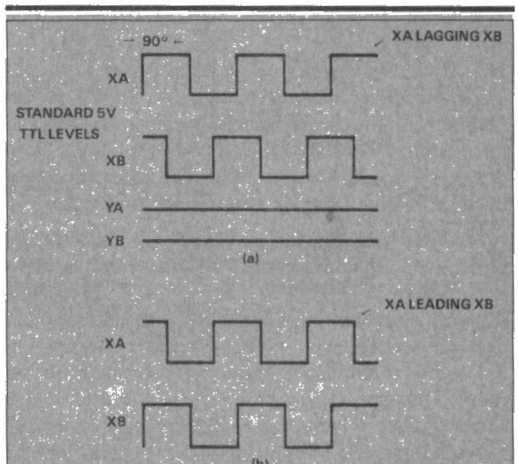


Fig. 3 Quadrature outputs from a normal mouse

HOW IT WORKS

The circuit diagram is shown in Fig. 2. It comprises a minimum system computer based around IC1, an 8031 single chip computer running at 4MHz. Its clock is taken from crystal XTAL1.

Address/data multiplexing is performed by IC2, controlling the access to the data in IC3, a 16K EPROM wherein lies the software required to generate the pulse trains explained below.

The joystick is fitted with four changeover microswitches. Depending on joystick position either none, one or two neighbouring switches are activated.

These four switches are polled in turn by part lines P1.4 through P1.7 of IC1, and when a switch press is detected the relevant pulse trains (see below) are produced on P1.0 through P1.3.

Let us take a closer look at the signals generated by the electronics of a normal mouse. The final output consists of two trains of square-wave pulses for the x and y direction movements of the mouse. The square-wave outputs are in quadrature (90 degrees out of phase) with each other and the phase relationship – leading or lagging – determines the direction of movement. No movement produces no output.

For example, if we move the mouse to the right the signals of Fig. 3a will be produced. Movement to the left will give the signals in Fig. 3b.

Similar phase related pulse trains are produced at YA and YB for a vertical movement of the mouse. For a diagonal movement, pulse trains are produced for both XA, XB and YA, YB outputs.

The number (or frequency) of pulses determines the amount of on-screen movement. With a normal mouse this is controlled only by the mouse ball rotating the optical encoded discs as shown in Fig. 1. The only way of altering the response on-screen relative to actual mouse movements is by software – either the operating system or the programme running.

With this circuit (Fig. 2) switch SW1 can be used to alter the basic frequency of the pulses (and so the response) to the user's taste in a very simple operation. Rocking the switch either way interrupts IC1 on pin 12 or 13 to increase or decrease (respectively) the frequency of IC1's resident timer.

Start up frequency is defined by address &003F of the EPROM (see Listing 1). This can be altered to suit your favourite pace and program.

Limits can be set to give a maximum or minimum speed that can be reached by rocking SW1. The EPROM address for minimum speed is &00B5, shown in Listing 1 with a data value of &00. Maximum speed limit is at address &00D0, shown as &FF.

Switches SW2 and SW3 have no interaction with the main circuit and simply ground the 'select' pins of the plug to the host computer – these 'select' switches are used in conjunction with a normal mouse to select software options and use will be described by individual programs.

```
0000 02 00 3E 02 00 B0 00 00
0008 00 00 00 C0 D0 02 00 60
0010 00 00 00 02 00 CB 00 00
0018 00 00 00 00 00 00 00 00
0020 00 00 00 00 00 00 00 00
0028 00 00 00 00 00 00 00 00
0030 00 00 00 00 00 00 00 00
0038 00 00 00 00 00 00 78 B0
0040 75 90 F0 75 8A 00 75 8C
0048 F0 75 89 61 D2 A8 D2 AA
0050 D2 A9 D2 AF D2 8C 80 FE
0058 80 FC 00 00 00 00 00 00
0060 00 00 88 8C 30 97 0B 30
0068 96 17 30 95 20 30 94 2A
0070 80 0C B2 90 7E 40 1E BE
0078 00 FC B2 91 01 6A D0 D0
0080 32 B2 91 7E 40 1E BE 00
0088 FC B2 90 01 6A B2 92 7E
0090 40 1E BE 00 FC B2 93 D0
0098 D0 32 B2 93 7E 40 1E BE
00A0 00 FC B2 92 D0 D0 32 FF
00A8 FF FF FF FF FF FF FF FF
00B0 C0 D0 B2 8C B8 FF 02 80
00B8 0D 08 7E 17 1E 7F FF 1F
00C0 BF 00 FC BE 00 F6 D0 D0
00C8 D2 8C 32 C0 D0 B2 8C B8
00D0 00 02 80 0D 18 7E 17 1E
00D8 7F FF 1F BF 00 FC BE 00
00E0 F6 D0 D0 D2 8C 32 00 00
00E8 00 00 00 00 00 00 00 00
```

Listing 1. Digital Mouse Hex Dump

PARTS LIST

RESISTORS (all ¼W 5% unless specified)	
R1	8k2
R2	4k7
CAPACITORS	
C1	10µ 16V tantalum
C2,3,4	100n ceramic
C5,6	33p
C7	47µ electrolytic
SEMICONDUCTORS	
IC1	8031
IC2	74373
IC3	2716
MISCELLANEOUS	
SW1	spdt (on)-off-(on) toggle
SW2,3	press-to-make button switch
J1	microswitch joystick
XTAL1	4MHz crystal
PCB. Ribbon cable. Nuts and bolts. Case.	

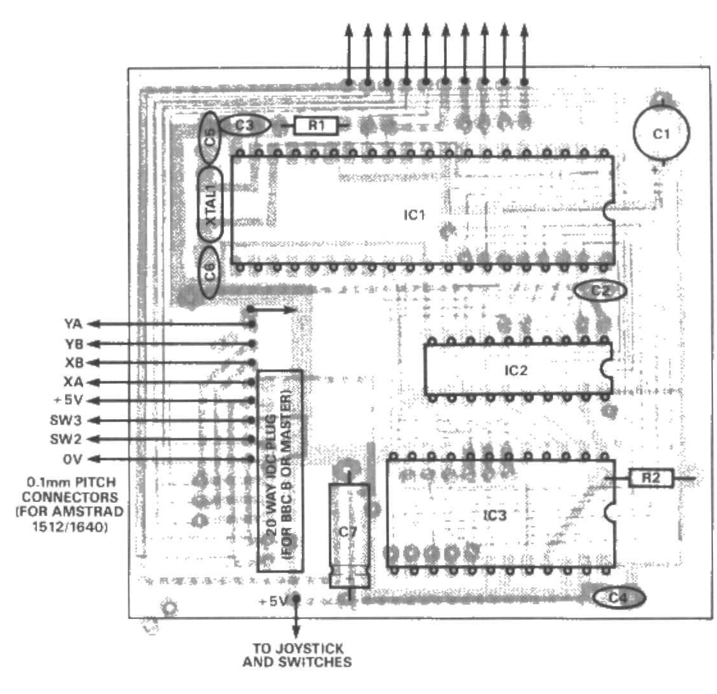


Fig. 4 Component overlay for the mouse-on-a-stick

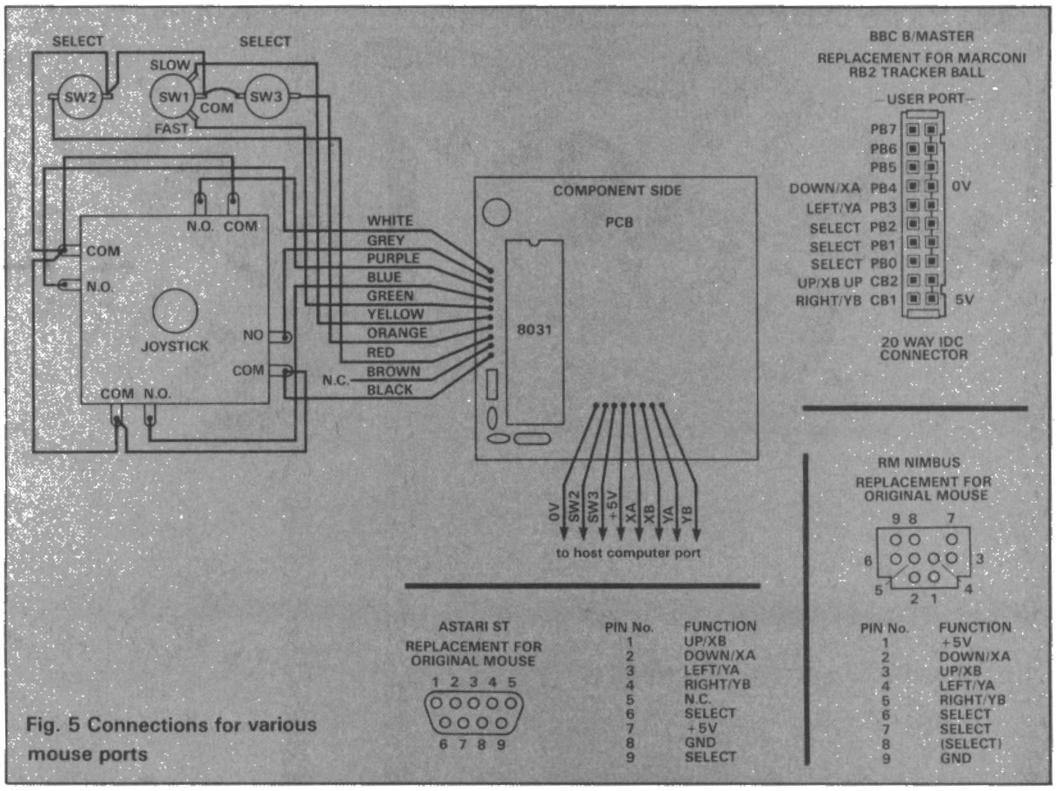


Fig. 5 Connections for various mouse ports

The prototype was constructed to interface with the Amstrad 1512 or 1640. Other computers can be easily accommodated by rewiring the interface lead to fit the inputs on your particular computer — various plugs are shown in Fig. 5. Refer to your user manual list of pins and wire up accordingly. A word of caution here, since plugging a non-manufacturer-recommended piece of equipment into your computer may well invalidate your guarantee. You have been warned!

Apart from that you should be able to get on and build yourself a joystick mouse into a case only slightly larger than the PCB. Desk space to spare and enough control to conquer the cosmos!

ETI

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ASP READERS' SERVICES (RO ET5/6),
Argus House, Boundary Way,
Hemel Hempstead, HP2 7ST

Please make cheques payable to ASP Ltd.
Overseas orders add £5 (Interspec) or £10 (Interbeeb) for airpost.
Access and Visa card holders can also place their order by phone on (0442) 41221
Allow 28 days for delivery.

Please supply ...... Interspecs (RO ET5) at £29.95 plus £1.95 p&p per order.
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Name ....

Address ....

....

....

Please debit my ACCESS/VISA card No. .... **to the sum of £** .... **Signed:** ....

INTERBEEB £49.95

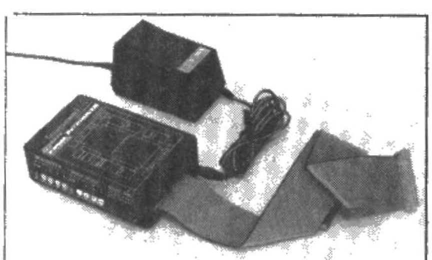
The Interbeeb unit connects to the BBC micro's 1MHz bus expansion connector and is supplied complete with its own power supply unit.

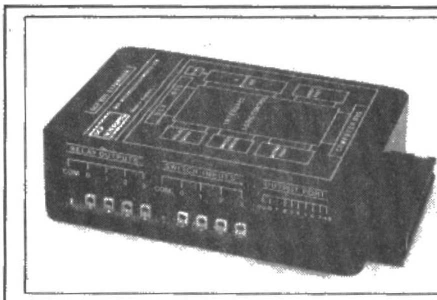
The interface unit is housed in a plastic case approx 4½x3x1in which contains the top quality double sided PCB and interface connectors.

- **8-bit input port**
- **8-bit output port**
- **four switch sensor inputs**
- **four relay-switched 12V 1A outputs**
- **eight channel multiplexed analogue to digital converter**
- **precision 2.5V reference**
- **external power supply**
- **15-way expansion bus**

All sections of the interface are memory mapped in the 1MHz expansion map for maximum ease of use and compatibility with existing peripherals.

The expansion bus provides all the data and address/control signals for the addition of further DCP modules or home-built devices. All the information required for using additional devices is included.





INTERSPEC £29.95

The Interspec unit plugs directly onto the expansion edge connector of the Spectrum to provide a full range of interfacing facilities.

The unit is housed in a plastic case approximately 4½x3x1in which contains the top quality double sided PCB and interface connections.

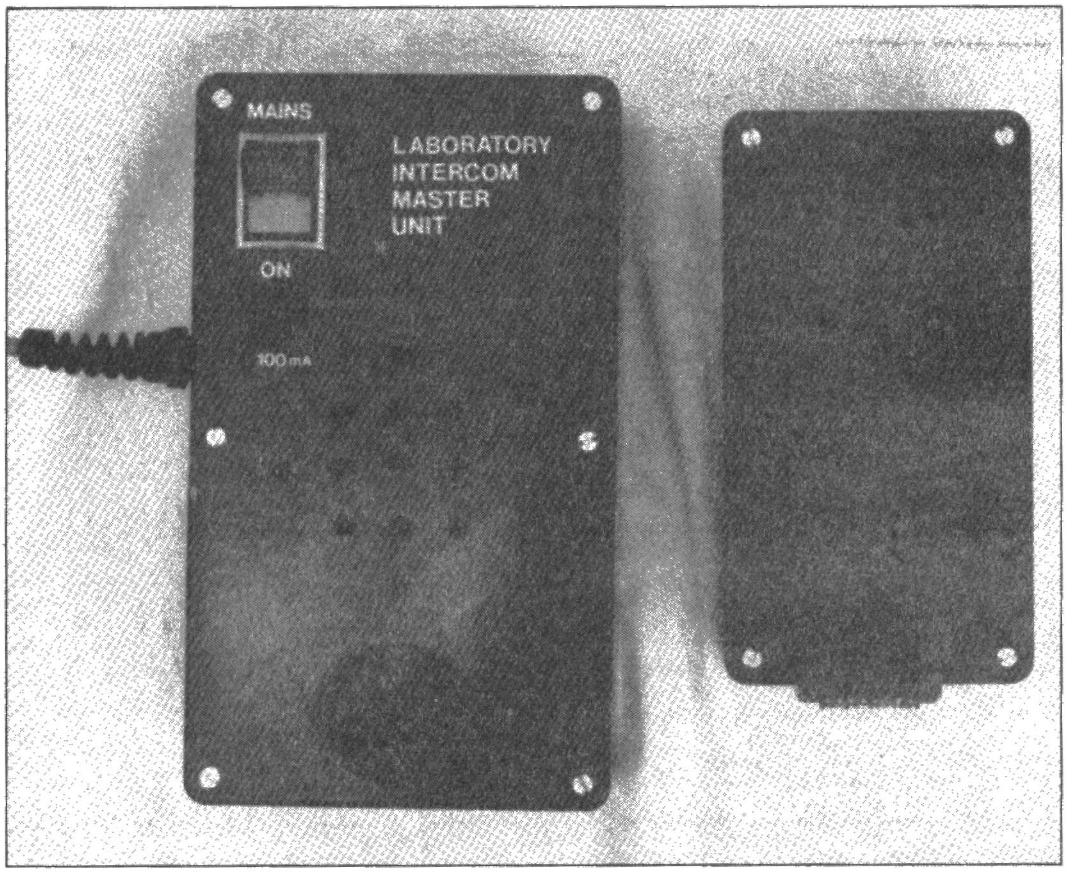
- **8-bit input port**
- **8-bit output port**
- **four switch sensor inputs**
- **four relay-switched 12V 1A outputs**
- **eight channel multiplexed analogue to digital converter**
- **15-way expansion bus**

All sections of the interface are I/O port mapped and designed for maximum compatibility with existing Spectrum peripherals. Power is supplied through the Spectrum edge connector.

The expansion bus provides all the data and address/control signals for the addition of further DCP modules or home-built devices. Connection is by multi-way PCB connector and all the information required for adding further devices is given.

SWITCHLESS INTERCOM SYSTEM

PROJECT



Messrs Lawson and Feeney can talk for hours on their multiway expandable intercom system

Ever since primitive man decided that he had something more to communicate to his fellows than the wants and warnings exchanged between less intelligent planetary cohabitants, he has striven to make communication possible over ever greater distances, between more and more individuals and against increasingly overwhelming problems of noise and surroundings.

Having now passed on from grunts and rock-banging to an age of digital-encoded, frequency-compressed, multiplexed-laser exchanges between land mobiles, geo-stationary satellites and micro-wave dishes, is there *really* a need for a new intercom design?

Well . . . yes there is. My particular problem existed in a hospital radio-pharmaceutical suite where a preparation lab is separated from the collection room by a changing area and double-glazed panels. This is where injections are prepared under sterile 'clean room' conditions. Because access is restricted, messages had to be communicated via the glass panels in a startling performance of shouting and improvised sign-language, not far advanced from the communications devised by the aforementioned primitive man.

A small commercial intercom was tried but since the clean-room operator spends most of his or her time at a bench with hands busy, the 'press to-talk' button (almost universal on small systems) was not convenient and communication tended to be one-way. Besides, the unit was always being left on and the batteries would flatten overnight.

So it was decided to attempt the unlikely — a simple two-way system free of 'press-to-talk' buttons.

Avoiding A Howler

People who have worked with microphones will have already spotted the major obstacle: a room with a live microphone feeding to a speaker in another room which contains another live microphone feeding a speaker in the first room, is an instant recipe for audio feedback — howl around.

Circumstances did not warrant excursions into possible technical solutions such as frequency shifters, adaptive filtering, line multiplexing and so on. It was judged that by reducing the acoustic coupling between speaker and microphone, sufficient sound levels could be achieved for useful working. If such a system would work, then another station was to be added in an adjacent clean-room.

Most cheap intercom systems rely on a single transducer, a loudspeaker, that functions as a microphone when suitably switched. This is mounted in a small plastic or metal box and the system gives a very poor tinny quality to the reproduced speech. A tinny sound will set up howling before useful sound levels could be reached, due to certain frequencies being reproduced more efficiently (mechanical resonances of the boxes, properties of the microphone and loudspeaker and so on) so that the whole system has sufficient gain to oscillate at these exaggerated frequencies at otherwise low levels.

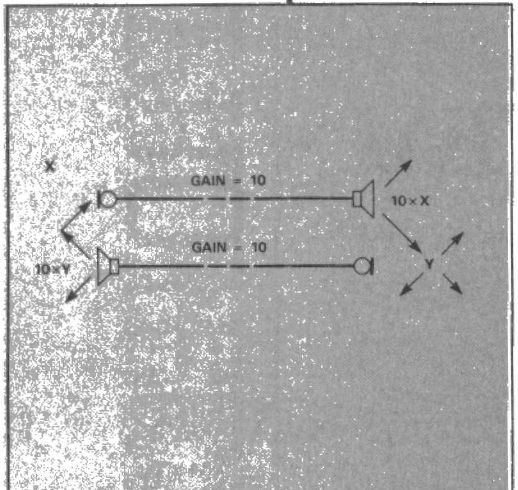


Fig. 1 Intercom feedback

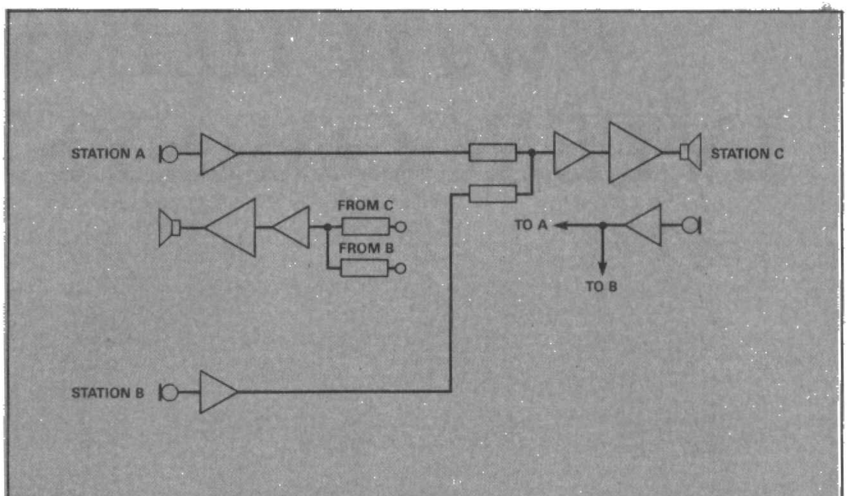


Fig. 2 Block system for three or more stations

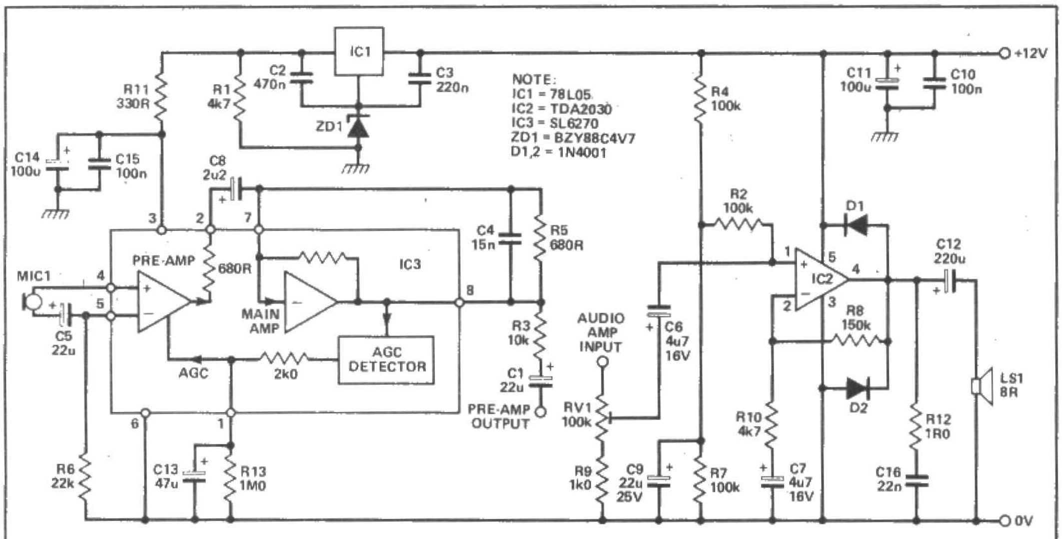


Fig. 3 Circuit diagram for a slave intercom station

PARTS LIST

MASTER UNIT

RESISTORS (all ¼W 5% except where stated)

R1,2,9	2M2
R3-8,11-13	100k
R10,14	4k7
R15,17,21	100k ½W
R16	10k
R18	680R
R19	150k ½W
R20	22k
R22	1k0
R23	4k7 ½W
R24	330R
R25	1R0 2.5W wirewound
R26	1M0
RV1	100k min cermet preset

CAPACITORS (all 16V radial electrolytic except where stated)

C1,4,5	220n tantalum
C2,3,9,14,20,24	100µ
C6,12	220n min layer
C7,11	470n min layer
C8	2200µ 25V axial
C10	22µ
C13	15n min layer
C15	4µ7 tantalum
C16	22µ 25V
C17	4µ7 63V
C18,23	100n min layer
C19	2µ 63V
C21	220µ
C22	47µ
C25	220n plastic

SEMICONDUCTORS

IC1	LM3900
IC2	7812
IC3	78L05
IC4	TDA2030
IC5	SL6270
ZD1	BZY88C4V7
BR1	1.6A bridge rectifier
D1,2	1N4001

MISCELLANEOUS

CONN1-4	2-screw PCB-mounting terminal blocks
FS1	100mA antisurge fuse
LS1	8R 0.5W min loudspeaker
MIC1	600R dynamic moving coil microphone (unidirectional)
SW1	mains rocker switch (incorporating neon)
T1	mains 12VA transformer 12V, 1A (split-bobbin type)

PCB. Large 1in grommet (for microphone). Black baize. Mounting bolts. Mains cable grommet or bush. Case.

HOW IT WORKS

The circuit of a slave intercom station is shown in Fig.3. It is a simple two amplifier design consisting of a microphone preamplifier and audio power amplifier.

The preamplifier is based around the SL6270 VOGAD microphone preamplifier, a voice operated gain adjusting device which uses an automatic gain control (AGC) circuit in its internal feedback circuitry to keep a constant amplitude. The AGC causes the gain to increase if the external noise is small and vice-versa — a useful feature for an intercom.

The only problem when using this device is that the maximum recommended operating voltage is 10V, whereas the minimum working voltage for IC2 is 12V. This means a voltage regulator is required to drop the 12V supply to around 9.5V for IC1.

However there are no commonly available 9.5V fixed voltage regulators so IC1, a 78L05, is used instead with a zener diode ZD1 in its common connection. The zener diode 'lifts' the reference voltage to the common connection by the zener voltage rating — to obtain a 9V supply, say, a 4V3 zener could be used (there is usually a small loss in the zener diode).

The gain for IC3 is set by R5 which provides negative feedback to the main amplifier of the chip. The gain is given by the following formula:

$$\text{Gain} = \frac{R_G \times 10{,}000}{(R_G + 10{,}000) \times 680}$$

The minimum value for RG is 680R, the value used in this application, which reduces the susceptibility to feedback due the AGC action to a minimum. This is required due to problems otherwise encountered with audio acoustic feedback inside the housing.

Anyone redesigning the unit as a switched intercom could remove RG.

Capacitor C5 provides a high frequency roll-off to help reduce 'howling' when feedback is produced. C13 and R13 set the rate of attack and decay of the AGC circuit, these are not critical and may need to be changed if a special microphone response is required.

R11, C14 and C15 help to decouple the supply to IC3 along with C11 and C10 for IC2. The amp output is fed into RV1 which acts as a volume control connecting to the non-inverting input of IC2.

R8 sets the gain and need not be changed. R12 and C16 form a zoebel network to give stability at high frequencies and D1 and D2 protect voltages larger than the rail.

For a simple two intercom project the master and slave units' inputs and outputs may be cross-linked. For a three station project a mixer circuit is required to isolate the stations properly.

A typical mixer circuit is shown in Fig. 4, based on the LM3900 quad Norton op-amp. This IC has current mirror inputs which allow single supply operation and around 95% of the supply voltage swing on the outputs, however the maximum input voltage is 0.6V. RG controls the gain which is set to unity for this project. If the signal needs a boost at this stage RG may be increased by make sure the output load due to cable resistance is not too high.

The two input resistors mix the signals from stations A and B and feed the output to the third station C.

The complete mixer circuit is shown in Fig. 5. Another station may be provided for by using the fourth op-amp on the 3900, with three resistors on each input instead of two.

The power supply is a standard 12V 1A type but make sure the transformer is of the split-bobbin type as the mains earth is not used and these transformers provide more isolation between primary and secondary coils. Use an anti-surge fuse in the live connection.

It became obvious that a live microphone system would have to have a fairly flat frequency response and that the penalty paid for the absence of switches would be a sound level somewhat less than can be used in a switched one-way system.

To produce a resonance-free frequency response it is important to choose better quality devices than would otherwise be selected. Various microphones and loudspeakers were tried before arriving at the present components. The amplifier was developed as a two chip design, a microphone matching pre-amp and a small hi-fi amplifier chip. This also permits the mixing needed for a larger system.

A directional microphone is a must. Looking at Fig. 1 a sound is produced at X and the mic + amp + speaker chain has a gain of ten. Some of this signal will find its way to the adjacent microphone and if this is any more than a tenth of the speaker level, then sufficient sound will be produced by the second channel to reinforce X and the system rapidly builds up to a teeth-grating howl. This is a much simplified explanation but illustrates why the microphones should not be responsive to sounds around or behind them and why mounting speaker and microphone via sound-insulating mounts is necessary.

Calling At The Next Station

With a system working, it is surprisingly easy to extend it to a large number of stations. A simple distribution system feeds each microphone to all loudspeakers except its own. For this, each amplifier is preceded by a basic mixer that accepts signals from various microphones (see Fig. 2).

By having each pre-amp next to its own microphone, the signals involved are then large enough to permit removing the mixer/distributor to a remote position or, for best economy of wiring, in a 'master' unit that comprises a power supply, the mixers and a single intercom station.

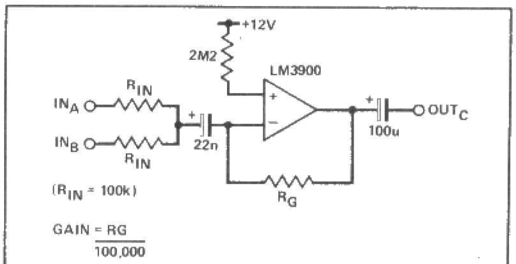


Fig. 4 Signal mixer using LM3900 Norton op-amp

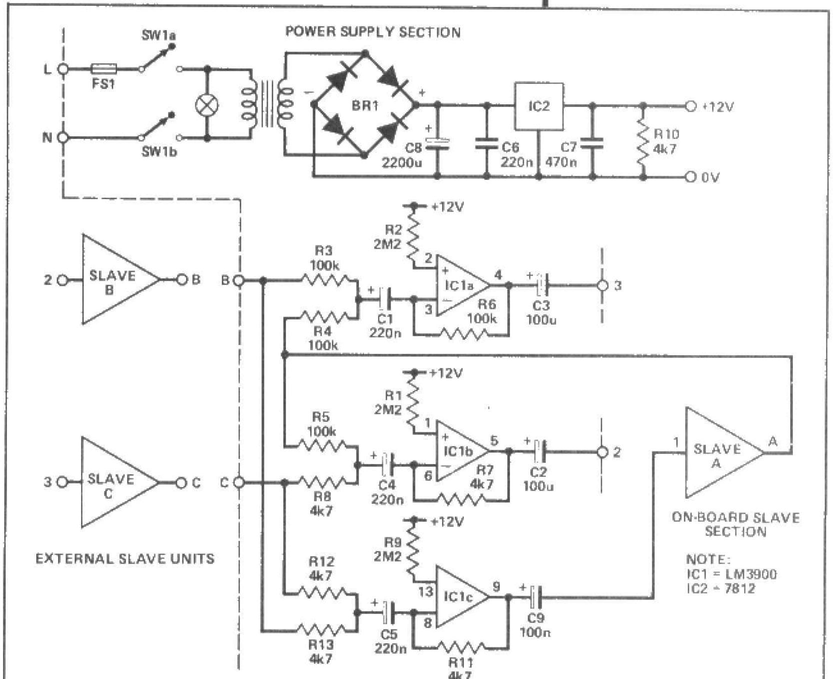


Fig. 5 Circuit diagram for the master station

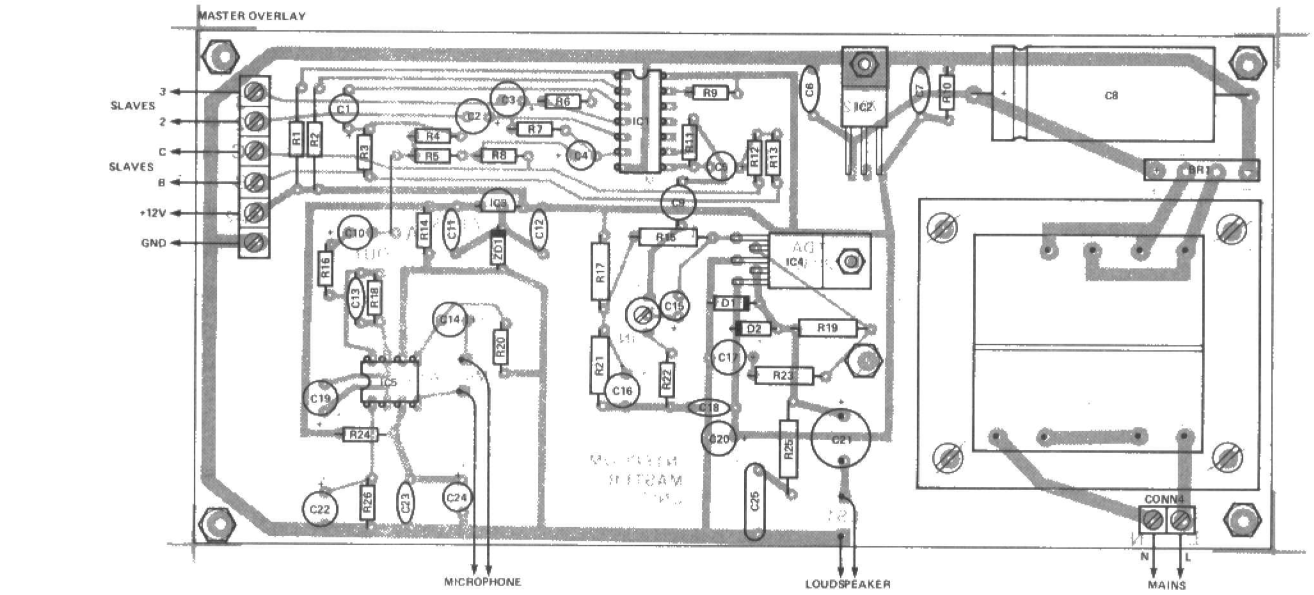


Fig. 6 Component overlay for the master station

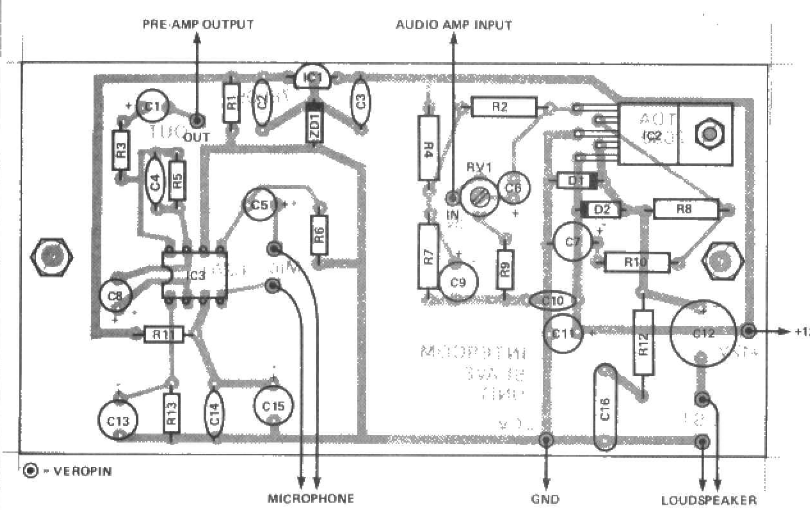


Fig. 7 Component overlay for a slave station

PROJECT

Construction

The overlay for the master unit is shown in Fig. 6 and the slave in Fig. 7.

When assembling the PCBs mount the flattest components first and build up according to height of components to make soldering much easier. IC sockets are not a necessity but are a useful addition in case of faults on the board. During testing it allows the supply to be checked before the ICs are placed in their sockets.

The TDA2030 and 7812 are mounted on the circuit board with a 4BA nut and bolt. Finish construction with the transformer, making sure it is secured with mounting bolts.

Make sure the box for the master unit is deep enough to house the transformer. Wire up the mains switch and fuse and connect to the terminal block. The circuit should be tested at this point for the correct supply voltages throughout the board.

PARTS LIST

SLAVE UNIT

RESISTORS (all ¼W 5% except where stated)

R1	4k7
R2,4,7	100k (½W)
R3	10k
R5	680R
R6	22k
R8	150k (½W)
R9	1k0
R10	4k7 (½W)
R11	330R
R12	1R0 wirewound 2.5W
R13	1M0
RV1	100k min cermet preset

CAPACITORS

C1,5	22µ
C2	470n min layer
C3	220n min layer
C4	15n min layer
C6	4µ7 16V tantalum
C7	4µ7 63V
C8	2µ2 63V
C9	22µ 25V
C10,15	100n min layer
C11,14	100µ
C12	220µ
C13	47µ
C16	22µ plastic

SEMICONDUCTORS

IC1	78L05
IC2	TDA2030
IC3	SL6270
D1,2	1N4001

MISCELLANEOUS

CONN1	4-screw terminal block with solder terminals
LS1	8R 0.5W min loudspeaker
MIC1	600R dynamic moving coil unidirectional microphone

PCB. Large 1in grommet (for microphone). Black baize. Mounting bolts. Case. Veropins.

Next solder the speaker and microphone leads to the PCB pins and socket the ICs (if sockets have been used).

Special consideration has to be given to the acoustics in the box as explained earlier. Figure 8 shows the mounting arrangements of the speakers and microphones used in the prototype. PVC tape was used for blanking off the back of the speaker. Tape deadens the internal sound better than most methods that were tried. The foam sealing strip is the material used for draughtproofing doors and this isolates the speaker from the plastic lid.

The black baize is glued in place first to act as further damping and to give a pleasing aesthetic finish. The foam strip is then put down in a circle the size of the speaker.

The microphone is isolated from the plastic lid by the rubber grommet — the microphones recommended require a grommet diameter of about 27mm but smaller ones could be used by cutting out the internal rubber. When the microphone is glued inside the grommet (leave ½in out of the lid), black baize is glued on top of the microphone for aesthetic appeal.

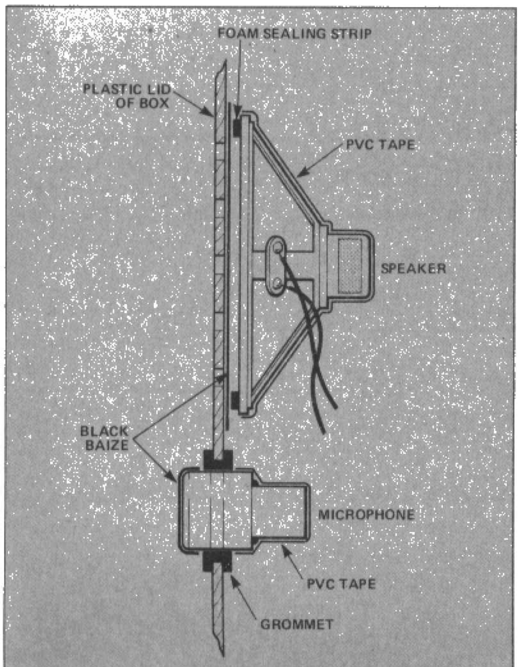


Fig. 8 Mounting the speakers and microphone

For the slave unit follow the same procedures but bear in mind the mounting method used for the boxes. The master unit is wall mounted with four screws underneath the PCB, with the PCB then swung into place on 4BA board spacers. The slave units are mounted using the same screws as the PCB.

Make sure to use 4-core-screened cable when wiring up the units from the mixer to the slaves. The terminal block used in the slaves is the solder-tag type where a hole is cut in the box to allow the terminal through.

The cases should be carefully cut so that everything fits in its proper place. Cutting instructions as used in the prototypes are shown in Fig. 9.

Testing

For the master unit test the power supply at the 7812 output with all ICs out of circuit and the microphone and loudspeaker disconnected. Next test the supply from the 78L05 which should be around 9 to 9.5V. If the supply is correct, turn off and socket the ICs.

Connect the loudspeaker and microphone and wire the two slave units to the terminal blocks.

On the slave units test the incoming supply with the ICs out of circuit and with the microphone and loudspeaker disconnected.

Next test the supply from the 78L05 (again around 9 to 9.5V), and if the supply is correct turn off and socket the ICs. Connect the loudspeaker and microphone.

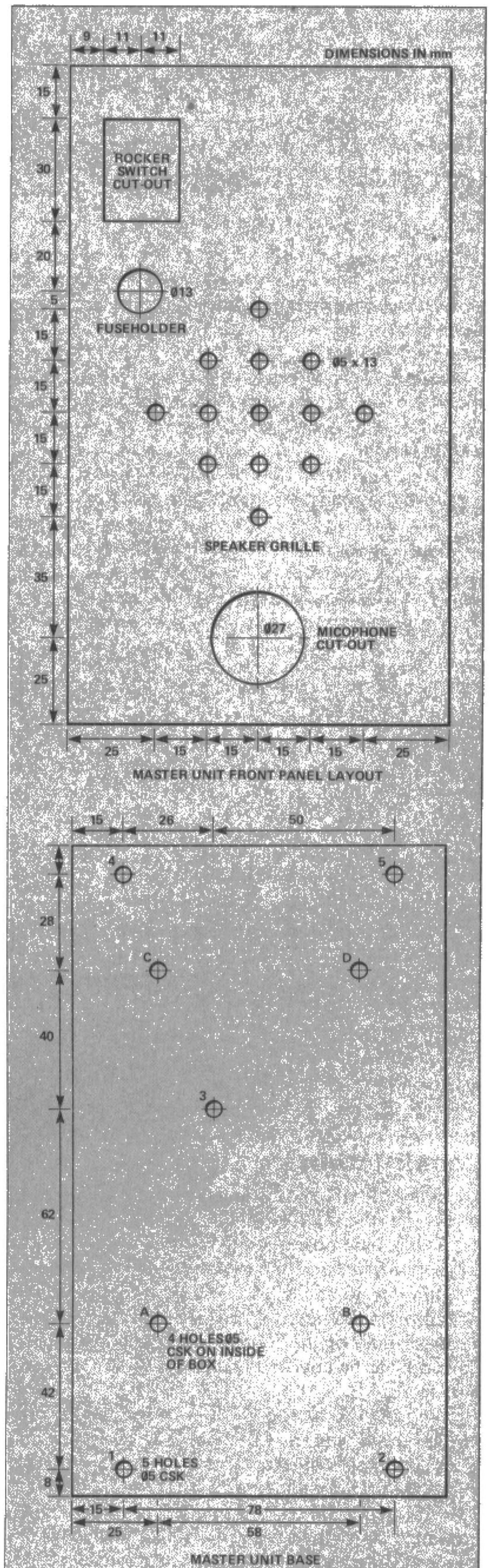


Fig. 9 Cutting instructions for the case

Setting Up

Space the two slave units a good distance apart (unless they are already installed) and switch on the supply. If any howling is heard, adjust the trimmers to prevent it.

Ask someone to speak into one microphone while you listen at another unit or use a portable radio to provide a sound source. Use your trimmer to give an acceptable volume at your end, then repeat the procedure for the other two units.

If the power mixer is being used, set all trimmers at highest value (unity gain).

If more volume is required turn each trimmer by the same amount until howling is reached and then back them off slightly.

Using The Intercom

The intercom was designed primarily to be switchless in operation. This has given rise to various feedback problems as previously discussed. To help prevent these getting any worse special precautions should be taken when installing the system.

Obviously the shorter the distance between units the better — if very long lengths of cable are required, the power mixer (outlined later) may provide a better performance.

All cable should be screened with the screen soldered to ground at either end.

Try not to position units too near a sound source as this will probably be fed through to the other units.

If the intercom is used in private rooms remember that the microphones are live all the time — it may be a good idea to design in a microphone muting switch for privacy purposes.

Be careful when routing the cables, making sure not to place near switching or power cables to avoid possible interference or signal degradation.

If you do not require the microphones to be 'live' all the time, by all means make the system a switched one. Increase the gain of the 6270 by either increasing the volume from R_G or removing it altogether.

Final Conclusions

The authors are honest enough to point out that this is not the ideal system for everyone by enumerating some of the drawbacks. There is no simple way to avoid at least two screened cables and a power supply pair to each station.

Sound levels are somewhat lower than usually found in intercom systems — but in practice will depend on exact construction and on just where they are used.

Because the system may have several stations, each contributing to the tendency to howl, the setting-up can be fairly critical for best levels.

In mitigation, the authors have produced a facility within bounds of cost and complexity as a solution to a problem not amenable to commercial equipment. Although many readers will not have an isolated clean-room as part of their lives, it is hoped the project will help solve similar problems and reduce muffled screams and arm-waving across the nation!

ACOUSTICS

As mentioned, the key to this project is the isolation of loudspeaker and microphone from an acoustic point of view.

Several arrangements were tried and constructors are free to experiment in this area. The ultimate separation is of course obtained by mounting the two in separate enclosures but this complicated the construction and is not ideal because each enclosure has its own resonances and the system is still liable to howl.

Short goose-neck stems were tried as microphone mounts but discounted because they actually moved the microphone further into the sound field of the loudspeaker (and incited colleagues to ribald comments about their appearance!).

Acoustic damping with rubber, foam sealing strip and PVC tape as indicated has soaked up sound waves from behind the loudspeaker. It is important to mount the microphone capsule at least ½in out of the enclosure to escape the residual sound within the enclosure.

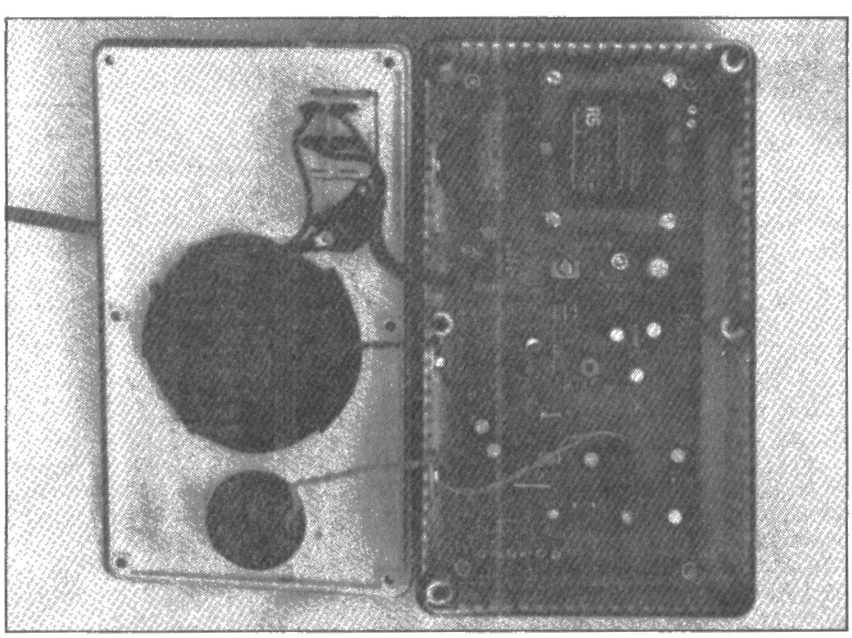
Areas for experiment include foam rubber tubes around the transducers, electret microphones, angled enclosures and so on.

BUYLINES

No problems should be encounted when trying to obtain components for this project. Various microphones were tried but the best was the unidirectional UF27 from Maplin.

The TDA 2030 and LM3900 are available from Maplin and Cirkit, but only Cirkit supply the SL6270. The bridge rectifier used was the SKB2/02L5A from Electromail. A suitable alternative bridge rectifier sold by Maplin is the BY164, which is also an in-line package.

The µA759 is only available from Electromail (telephone (0536) 204555). The transformer used was Electromail code 207-699.



The Power Mixer

The installation of a second system presented a new problem since the authors had to use wiring already in place in conduits in the walls. None of this wiring was screened so that any two wires formed a capacitor such that each microphone signal was fed straight back to its own loudspeaker. The capacitance between two adjacent rooms was measured at about 60nF giving almost total feedback at the high impedance amplifier.

The solution to this is to move the amplifiers so that only high level, low impedance levels are concerned, and 60 nF is not significant.

This is not too difficult with two stations but three or more require the mixer/distributer to be high level also, so that the circuit of Fig. 10 replaces the LM3900 mixer circuit.

The amplifiers are 759 devices that 'look like' 741 devices but have a beefy 300mA drive capacity, enough to drive over 700mW through an 8R loudspeaker. That might not seem much in these days of mega hi-fi but is uncomfortably loud in many circumstances!

The 759s are simply biased so they can be used from a single 12V supply. Gain is adjusted via RV1-3. If signal strengths are weak then increase the gain of the appropriate mixer op-amp.

A suitable component overlay is shown in Fig. 11.

To use the power mixer, the following wiring alterations need to be made to the slave units.

Connect the 6270 output directly to the TDA2030 input. Disconnect the output from the TDA2030 to loudspeaker and connect the loudspeaker to terminal 3, then re-connect the TDA2030 output to terminal 4.

Now terminal 3 is the output from the mixer and terminal 4 is the input to the mixer from the slave unit.

Alterations to the master unit are as follows:

Remove the LM3900, IC1. Remove the link from master unit PCB. Insert a PCB pin next to C10 and solder a lead from it to 'A' on the power mixer.

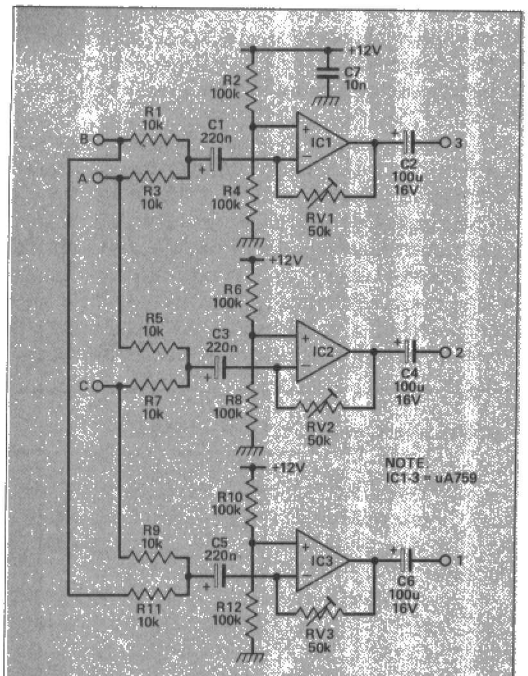


Fig. 10
Circuit diagram for the optional power mixer

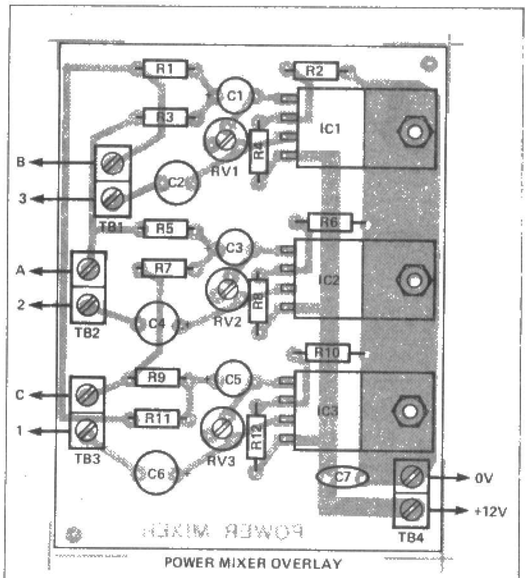


Fig. 11
Component overlay for the optional power mixer

Remove C9 and insert a PCB pin where C9's negative lead was. Solder a lead from this to '1' on the power mixer.

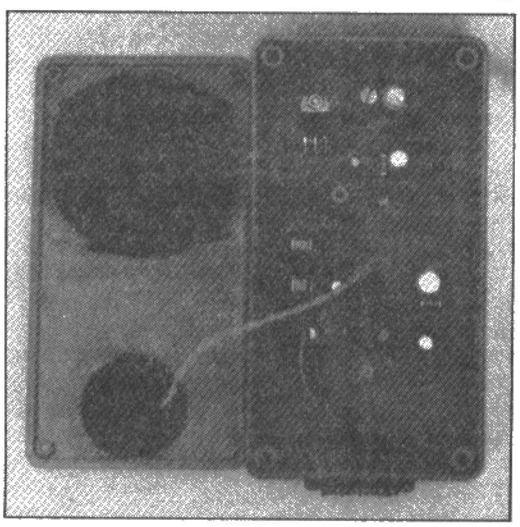
Mount the power mixer PCB on the side of the box and connect up B, C, 2 and 3 to corresponding terminals.

Connect +12V and Ground to power terminal block on power mixer.

PARTS LIST

POWER MIXER

RESISTORS (all ¼W ± carbon film)	
R1,3,5,7,9,11	10k
R2,4,6,8,10,12	100k
RV1-3	50k min cermet preset
CAPACITORS	
C1,3,5	220n tantalum
C2,4,6	100µ 16V electrolytic
C7	10n ceramic
SEMICONDUCTORS	
IC1-3	µA759
MISCELLANEOUS	
CONN1-4	2-screw terminal block PCB-mounting
PCB. Mounting screws, 4BA nuts and bolts for mounting 759s.	



REFLEX ACTION

MICRO MONITOR SPEAKERS

Jeff Macauley gets out the glue and reports on the ports of his bass reflex loudspeaker design

Last month I described in some detail the basic principles behind the design of reflex speakers. This month I've taken the opportunity to get my hands dirty and actually build one. There's nothing to beat the practical application of theory and so I present the following design which I hope will prove both instructive and a spur to like-minded readers to have a go!

Despite the fact that the theory behind speaker design is now fairly well explored, building a speaker will always throw up interesting challenges. For example this project started with running the characteristics of some woofers through the *Optibox* computer program I detailed last month.

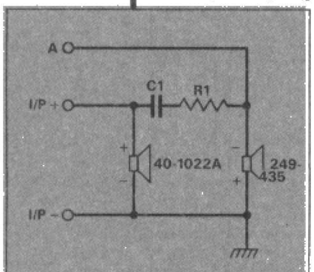


Fig. 1 The circuit diagram of the crossover

Although I live in a flat where space for audio equipment is somewhat limited I like my fair share of bass so I'm always on the lookout for small systems with good low end extension. I came across the woofer used in this project in the Tandy catalogue. Although it's only four inches in diameter and is primarily intended for in-car use its characteristics make it ideal for a micro monitor application. The unit, Tandy catalogue number 40-1022A, has the following specifications.

Resonant frequency	fo	=55Hz
Total Q	Qts	=0.35
Equivalent volume	Vas	=0.23 cu ft

If we plug these values into the equations given last month we come up with the following information. Optimum volume Vo = 0.169 cubic feet, cutoff frequency 62Hz!! I've seen enclosures of over a cubic foot which cannot reach that low.

This obviously merited further investigation so I went down to my local Tandy store and bought a pair. I then embarked on the process of testing them to make sure the parameters were as stated. I was not disappointed. Both units measured up very closely to the published spec and in addition the enclosed data sheet promised a very smooth midrange response and a top end rolloff of about 5kHz.

There were only two flies in the ointment. Firstly the quoted sensitivity of 84dB/W is on the low side as is the published power rating of 10W rms. However I have been driving mine at neighbour-disturbing levels from my 35W/ch amp for over a month with no signs of distress to the drivers. My initial reservations as to the spl obtainable and the woofer's durability have proved unfounded.

I finally chose, after some deliberation, a cabinet volume of 0.166 cu ft. The reason for the slight change in volume was rounding the dimensions to whole numbers: 3dB cut off at 64Hz, ripple 0.1dB. The final external dimensions of the enclosure are 9.5in by 5.75in by 9in (h × w × d). A cabinet this small has the advantage that the panels are more rigid and hence less liable to resonate and colour the sound. I eschewed the normal .75in thick high density chip board for the more cosmetically pleasing white melamine covered variety.

Any constructor that cares to can choose the high density stuff in preference as long as the internal dimensions are retained. An 8ft length of 9in wide board will amply provide two cabinets for less than £7. Unless one actually enjoys woodwork or has masochistic tendencies it is worth getting your local timber yard to cut the board for you. It is absolutely essential to have the cuts made accurately (this will prevent the possibility of the air turning blue as you fit the case together).

To complement the woofer a high quality tweeter is required. As with woofers there is a large variety on the market. I finally chose a Philips polycarbonate dome type which is available from Electromail, catalogue number 249-435. The main reason for this choice is that I have had wide experience of this unit and know its strengths. It has a good flat response from3-20kHz, amplitude variations being kept to within a couple of dB. It is also very sensitive and capable of wide dispersion.

Considering Crossovers

Having decided on our drivers and the cabinet size, the next task is to decide on a crossover. To understand how well or dismally a crossover works it is first necessary to consider the tasks it is required to perform.

The obvious function is to divide the incoming signal into the high and low components. Unfortunately this is complicated by the fact that the speaker units don't present pure resistive loads. Another complication is that back emf is generated by the units as they operate and this behaviour is both amplitude and frequency dependent. Add to this a compensation for response deviations in the driver and you end up with a very complex network that evades straightforward analysis.

The other alternative is to accept that some variations are inevitable and minimise these by choosing drivers that are inherently flat in their operating range. A simple network can then be designed which gives good results. Luckily both the units chosen here are fairly flat (to within a couple of dB of their operating range) and therefore the simple approach can pay dividends.

Getting back to the immediate problem of designing a crossover for the speaker in hand I made the following observations. Firstly the frequency response of the woofer is flat within ±2dB and has a well damped hf rolloff, 3dB down at 5.5kHz. Secondly the tweeter is far more efficient than the woofer and has a well controlled 'bass' resonance at 2kHz. It is also sensibly flat between 5 and 20kHz.

It followed from these observations that the easiest solution was to leave the woofer response as it is and precede the tweeter with a first order network. It means an asymmetric crossover and implies a careful choice of turnover frequency to avoid response

PROJECT

anomalities. This was the design path chosen and the resulting circuit is shown in Fig. 1. The tweeter is several dB 'hotter' than the woofer which means its signal must be attenuated. This is the function of R1. C1 is chosen to produce a −3dB point at 5.5kHz in conjunction with the series resistance of R1 and the tweeter's impedance. Note that the tweeter is phase inverted with respect to the woofer due to the drive signal phase differences between them.

Despite its simplicity this network does the job it was intended to do and has a couple of advantages compared to more conventional designs. It's extremely easy to drive, the impedance doesn't fall below 7R. It uses no coils and because of the high value of R1 a normal value polyester cap can be used for coupling.

Even so, I always find the advantages of an active crossover boost performance further, and a suitable design for the Micro Monitors will follow next month.

Construction

The required wood cuts are shown in Fig. 2. While dimensions here are in millimetres, the *Parts List* quotes in inches since many timber yards have yet to find their way in to the '80s!

Mark out the panels for the drivers and port. The tuning duct is a length of plastic piping with an internal diameter of 1.25in (31mm) which has an overall external diameter of 1.5in (37.5mm). The internal area is thus $3.14r^2 = 1.22\ in^2$. Plugging this value into the port equation gives a length of 4.1in, 104mm. The piping used is available at plumbers supply merchants and a metre length cost me all of 84p.

Note that the port is mounted on the rear baffle. This is mainly because a small frontal area is required to give good stereo imaging. It has no adverse effects on the sound because at the low frequencies at which the port operates the speaker's radiation pattern is omnidirectional anyway. Cutting the pipe accurately is more difficult. A mitre box is almost essential. If you don't have one the easiest way is to wrap a piece of card tightly around the tube to obtain a straight edge which can be slid up and down. Mark off the length you need by drawing around the card. Using a hacksaw carefully make a cut around the circumference of the tube before cutting it through.

Having got to this stage, final assembly of the enclosures can commence. Everyone is entitled to their own foibles when it comes to adhesives. Mine is to use a contact adhesive *Thixofix* which is widely

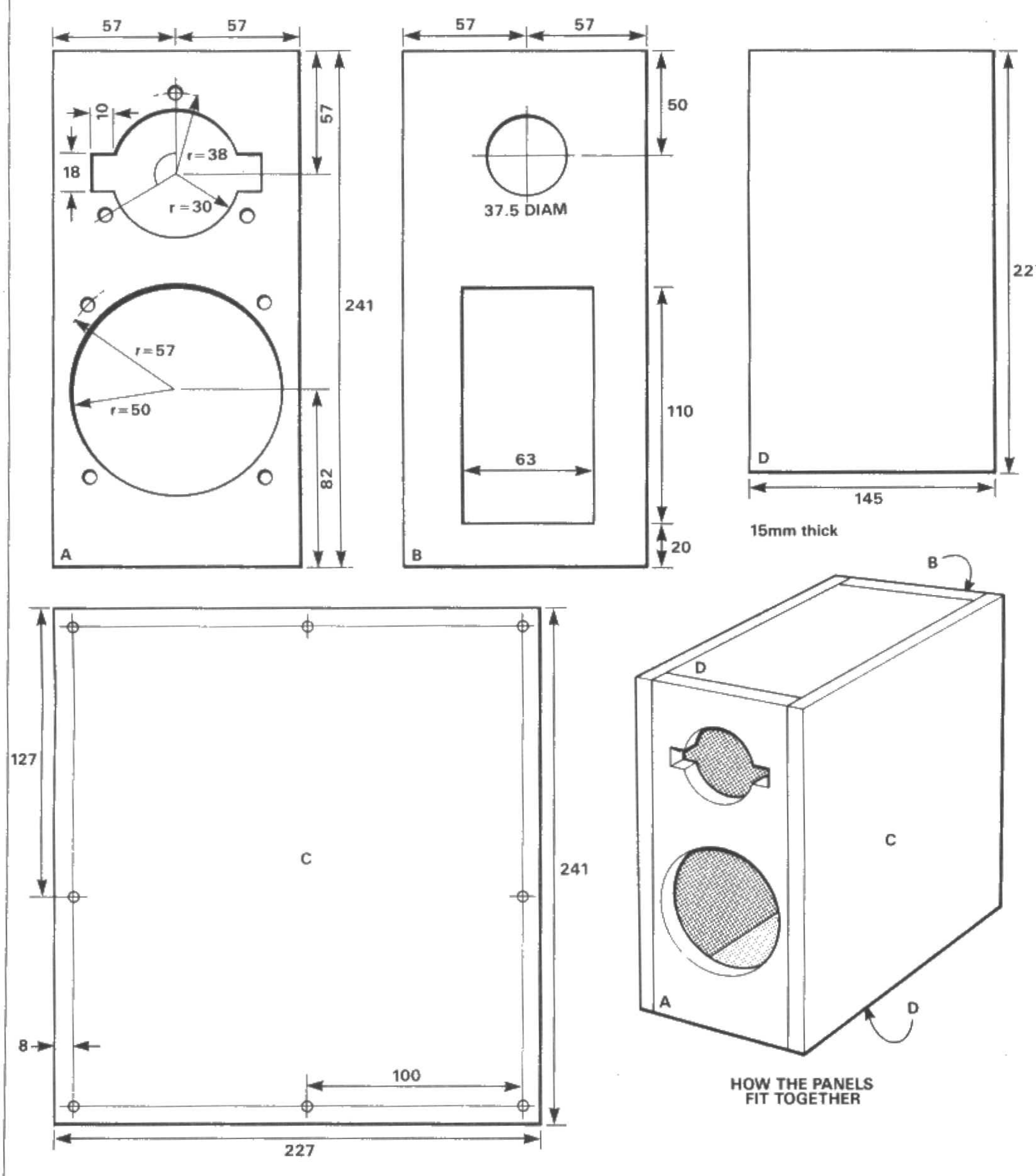


Fig. 2 Cutting diagram for a single Micro Monitor and enclosure construction

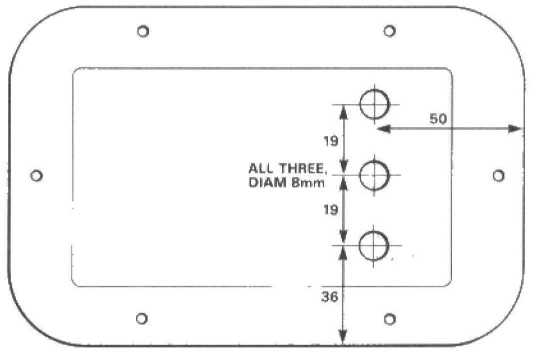


Fig. 3 Recess dish drilling detail

available. One smears the surfaces to be joined with glue, then leaves them for ten minutes or so for the glue to cure. The panels can then be slid together and positioned precisely. Firm pressure then completes the bond.

Whatever adhesive is used it is as well to use it liberally as this will stop air leaks.

Once the case has been glued it needs to be screwed together. I used 1in no8 self tappers for this job. Drill pilot holes ⅛in (3mm) diameter to take the screws and countersink these. I have detailed the screw positions on Fig. 2. These should be adhered to.

Having built the case, attention can be turned to sealing it to prevent air leaks. This is most easily done with 'polyfiller'. Mix it up into a stiff paste and work it along the panel seams with your finger. Wipe the excess away with a damp cloth. The hole for the port is best cut with a hole cutter. A nest of these, suitable for cutting various diameter holes between one and two inches, can be obtained from your local tool shop for a few pounds. These are a useful addition to anyone's tool box. To fit the port into position I used araldite rapid. The other apertures are larger and are best tackled with a jigsaw or jigsaw attachment. If this is not available then a coping saw can be used.

Reflex enclosures cannot be stuffed with sound absorbants like other forms of enclosure. To do so will wreck the carefully calculated relationship between driver and enclosure resonance. For this reason the interior is left unstuffed.

One nuisance with the woofer is that the mounting holes are slightly recessed from the frame. This means that in order to seal the driver against air leaks, further action must be taken. The best idea is probably to stick some self adhesive draught excluder tape around the hole before mounting the driver. A less strenuous alternative used on the prototypes was to seal around both units with polyfiller after mounting them. The tweeter hole should be similarly treated.

Before fitting the drivers into position it is as well to solder the leads to the terminals. Because of the simplicity of the crossover unit this is simply wired across two of the terminals as shown in Fig. 4 drawing. Nothing elaborate is required here — I used 12in lengths of 5A twin speaker cable from Halfords. This has the advantage that one of the conductors is identified with a white strip along the insulation thus allowing easy phasing of the units.

For connections to the drivers I used three 4mm banana sockets/channel. If these are connected as shown in the schematic then upgrading the speaker to active operation becomes easy. The drilling for the rear access dish is shown in Fig. 3.

Final assembly consists of mounting and wiring the crossover and terminals then screwing the rear panel into position. Remember to seal around the back plate with more polyfiller to ensure airtightness.

At this stage all that remains is to test the unit out and finish the cabinet as desired. If you've never used melamine faced board before you will find that edging strip is available to cover the bare chipboard. This is first trimmed to the correct width and applied to the edge by ironing over it. The backing adhesive bonds with the heat.

In conclusion this pair of speakers can be built for a total expenditure of less than £40 the pair. They have a better bass response for their size than any others I have heard. What's more important is that this response is nicely damped with no audible ripple in the response. They also offer a good transient response and a 3-dimensional stereo image. They can also be simply and fairly cheaply upgraded to active use as detailed in a forthcoming article.

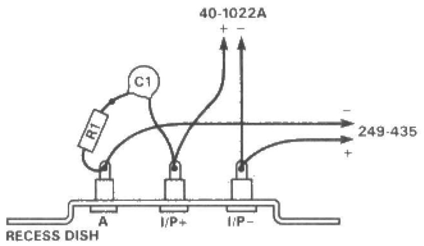


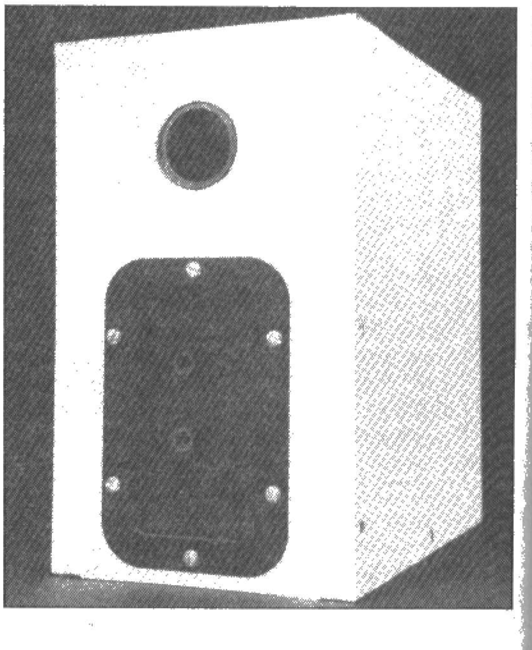
Fig. 4 Wiring diagram

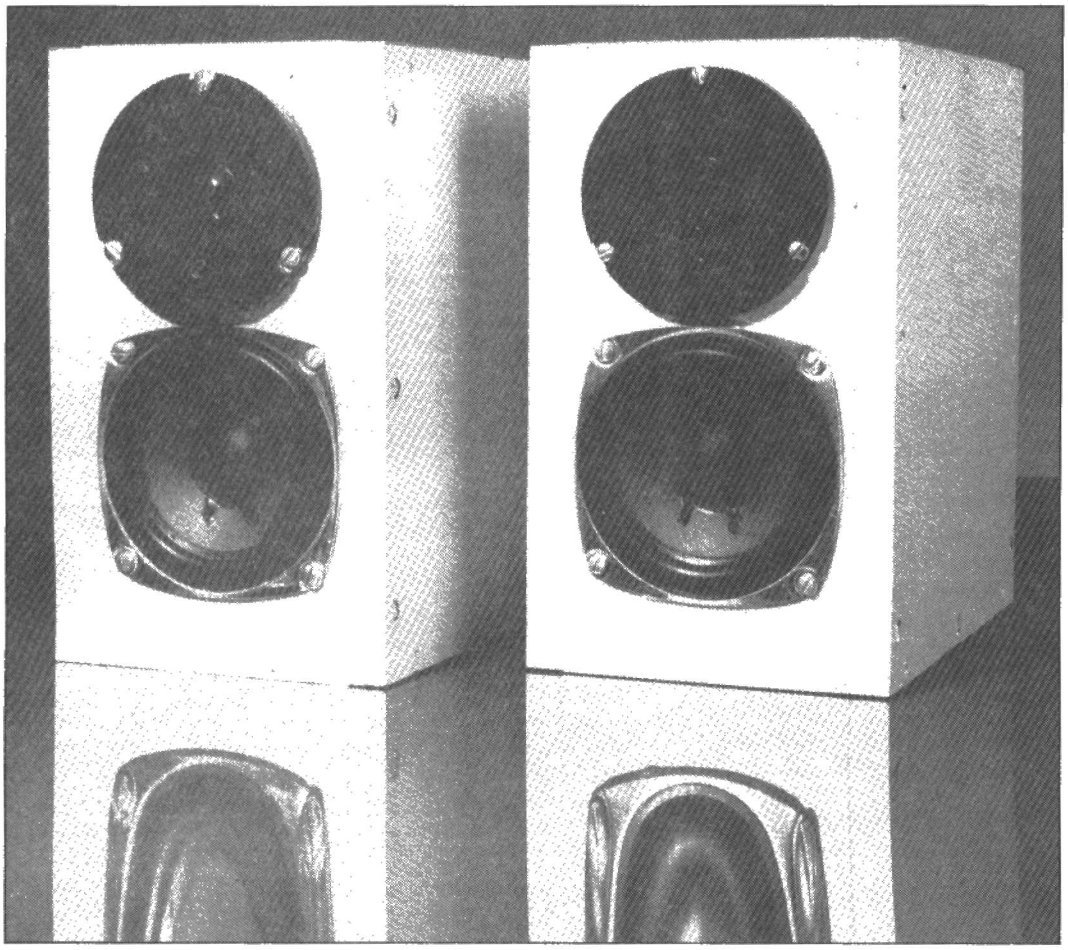
PARTS LIST

ONE CHANNEL ONLY

R1	47R 1W wirewound
C1	470n 100V polyester wkg
Woofer	Tandy 40-1022A
Tweeter	Electromail 249-435
Recess Dish	Maplin FS34M
Wood	15mm thick Melamine covered chipboard 2-off 9½ × 4½in 2-off 9½ × 9in 2-off 4½ × 7¾in

104mm length of 1¼ in id plastic tubing. 3-off 4mm panel mounting sockets.





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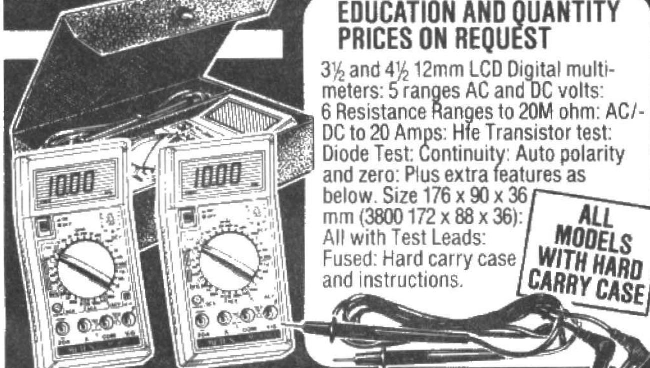

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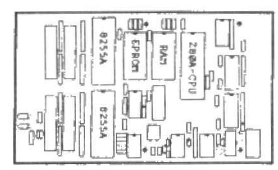

SINGLE BOARD COMPUTER "SBC-1"

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A star feature is that no special or custom chips (ie PALs, ULAs, ASICs etc) are used — and thus there are no secrets. The Z80A is the fastest and best established of all the 8-bit microprocessors — possibly the cheapest too!

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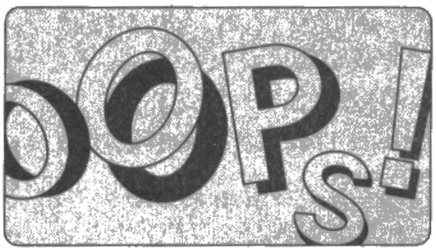


Peak Programme Meter (October 1988)
In Fig. 4 D21 and D22 should be swapped, C13 should be 100u and have no connection to IC12. The anodes of D15 and D17 should be connected to the input of IC12 (negative of C11).

On Fig. 7 the capacitors from top to bottom should be labelled C10,13,9,11,12.

On the Parts List R15,36=51k. R44,45,46, 49,50,51=2k7.

Chronoscope (November 1988)
In the overlay diagram for the counter PCB (Fig. 3) the polarity of C12 is shown the wrong way around. SW1a-d is shown as SW1-4. In Fig. 4 the cathodes of LED 8 and 9 are the righthand and lefthand pads respectively. The cathodes for LED 6,7 are marked as the wrong pin. In the text section on Battery Operation, Q1 should read T1. In Fig. 5 SW2 is incorrectly labelled SW5.

Doppler Speed Gun (December 1988)
In Fig. 2 the labelling of pins 7 and 4 of IC2 are transposed. IC10a Pin 1 and IC9c Pin 10 should connect together and not to the 5V rail. The positive terminal of C3 should connect to the junction of R2/R3. Pin 7 of IC2 should connect to the 12V rail and not to Pin 6/R1. So the pin labelling of CONN1 runs left-right on the overlay diagram, the corresponding labelling in Fig. 2 should be 3-1-2, reading downwards. Fig. 4 is correct in all respects except for the orientation of Q2 for which the c and e labels should be transposed. In addition the extra switch to be seen in the photograph of the prototype is a hangover from a previous incarnation. Just ignore it!

Burglar Buster (December 1988)
The foil part of the component overlay for the basic alarm (Fig. 1) was printed the wrong way around. It should be rotated through 180° as in Fig. 5.

Rev-Rider (January 1989)
In the Parts List RV2 is incorrectly given at 33k. It should be 22k as in the circuit diagram. A 'blob' went missing from the circuit diagram. RV2, R7, R4, C1 and D3 should all be connected.

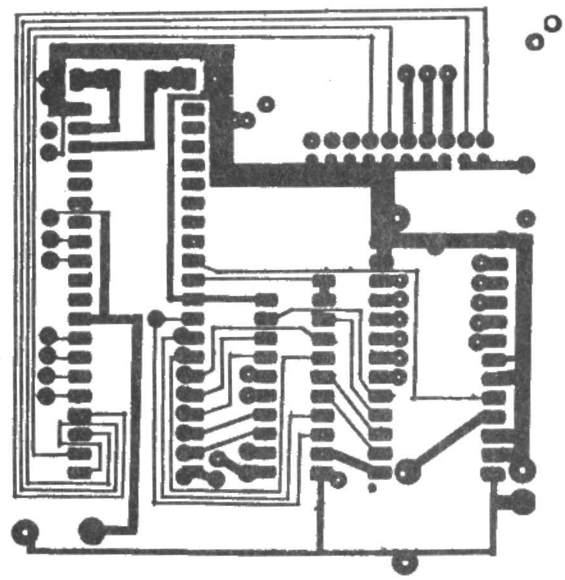
In-car Power Supply (January 1989)
Fig. 3 shows the front view of the 317 regulator with the pin-outs reversed. The photograph, circuit and overlays are all correct showing the ledge at the front of the device.

Audio Design MOSFET Amp (May 1989)
For home constructors of the power amp PCB (Fig. 8), the copper area connecting the negative of C7, C14 and R20 is a 0V #2 connection and should be linked to the 0V #2 copper area at the junction of C16 and C18+. Hart's kit PCB has a ground plane and no mod is necessary. Note that the preset at the bottom right of Fig. 8 takes the place of an external RV3 rheostat when bench testing and is not normally required.

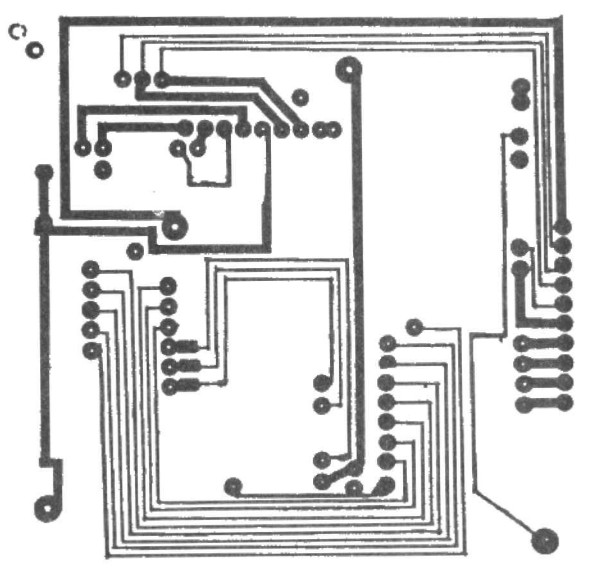
Bench Power Supply (May 1989)
In the Parts List, Q3,4 should be BC237 not BC307. The value in the circuit diagram is correct.

How To MIDI A Piano (June 1989)
In Fig. 5 the connection from pin 19 of IC8 ($\overline{MREQ}$) should go to pin 12 of IC7a, not pin 13 as shown. The component overlay is correct.

PCB FOIL PATTERNS

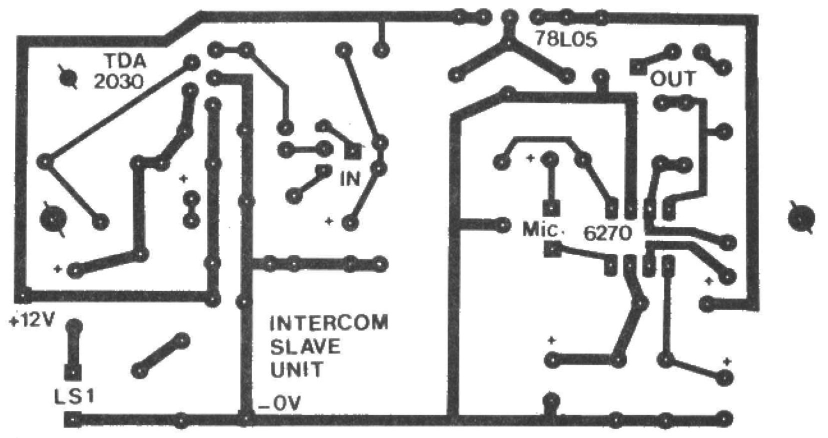


The digital joystick-to-mouse conversion foil pattern solderside

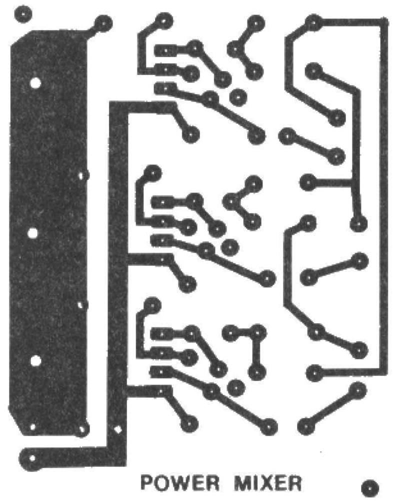


→ Added to ETI June 1989

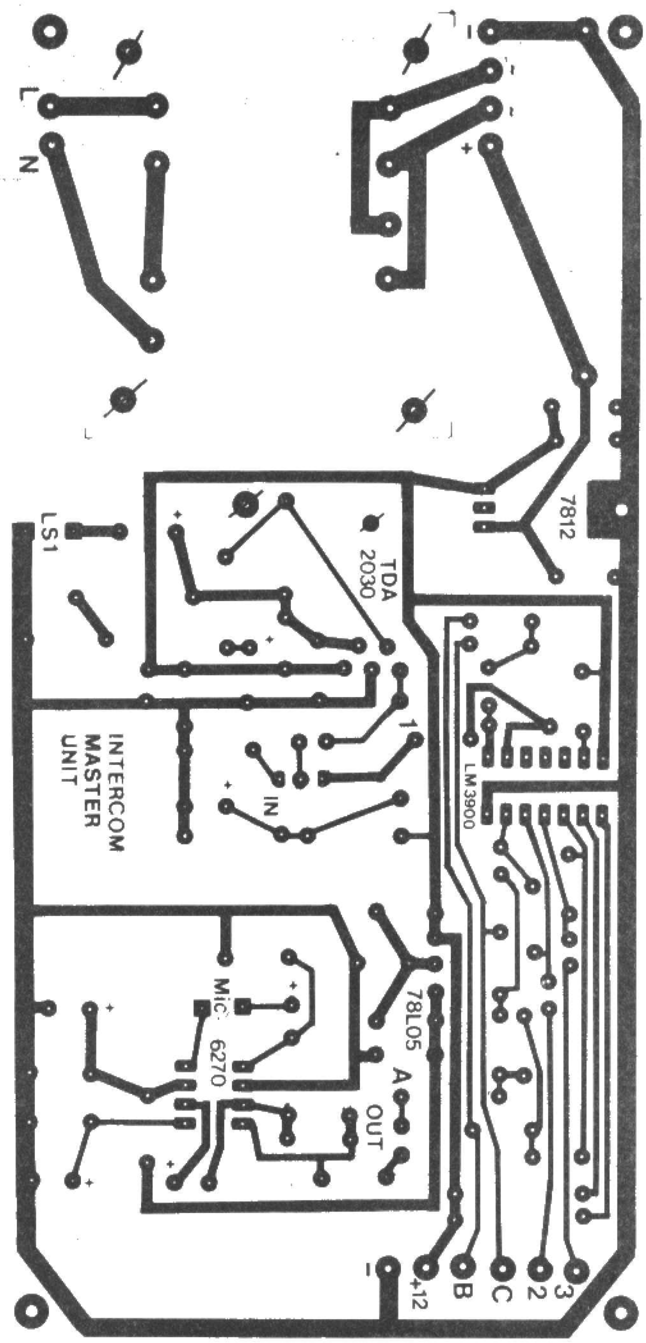
The digital joystick-to-mouse conversion foil pattern topside



The intercom slave station foil pattern



The intercom power mixer foil pattern



The intercom master station foil pattern

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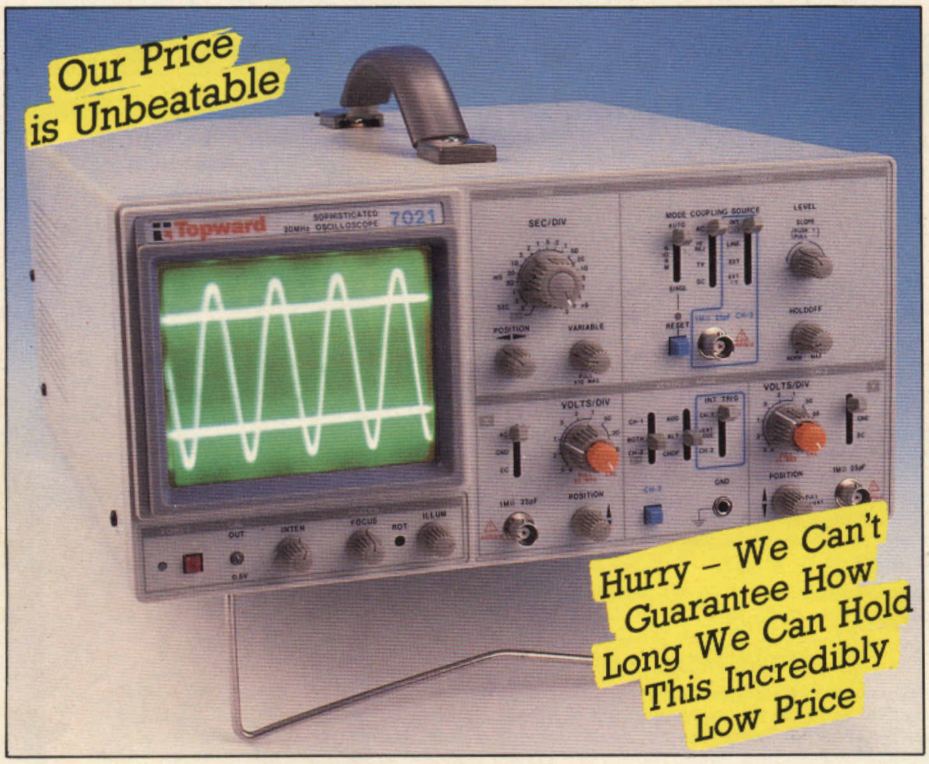

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